VOL. 14 No. 12, pp. 345-376 JUNE 1953

Philips Technical Review

DEALING WITH TECHNICAL PROBLEMS
RELATING TO THE PRODUCTS, PROCESSES AND INVESTIGATIONS OF
THE PHILIPS INDUSTRIES

EDITED BY THE RESEARCH LABORATORY OF N.V. PHILIPS' GLOEILAMPENFABRIEKEN, EINDHOVEN, NETHERLANDS

AN EXPERIMENTAL 100 kW TELEVISION OUTPUT STAGE

by D. ZAAYER. 621.397.645.026.445

A laboratory model of a television output stage has been completed for frequencies up to 68 Mc/s, having a resonance curve suitable for a 6 Mc/s band-width and delivering an output of 100 kW. Output power of this order provides good television reception for some distance beyond the optical horizon.

Television transmitting stations are usually established in the centre, or on the fringe, of densely populated areas. If such an area, as seen from the transmitting aerial, lies largely within the optical horizon, a picture and sound transmitter of a few kilowatts is large enough to cover the whole of it. For a more extensive area, however, considerably more power is needed to give good reception some distance beyond the horizon.

Another reason for the use of powerful transmitters is found in the fact that in many cases the receiving aerials employed are not very effective, either because the owner prefers an indoor aerial to one on the roof, or because the aerial has to be suitable for reception from different directions and at different wavelengths, and this leads to a rather unsatisfactory compromise.

Accordingly, it will be useful to look into the problem of generating considerable power at the frequencies on which television transmitters operate. As far as such frequencies are concerned, the first consideration in any research will be the lowest band of frequencies that has been made available for television: this is from 41 to 68 Mc/s.

A difficulty arises in that the band-width of the video transmitters has to be quite considerable, e.g. 6 Mc/s for a television system of 625 lines and 25 frames per second.

Under this system each TV frame consists of $^4/_3 \times 625^2 \approx$ 520.000 picture elements. (The width of the picture is $^4/_3 \times$ the height.) Assuming that the lines are made up of alternate white and black image elements, the time taken to scan two adjacent elements corresponds to one cycle of the video signal. The whole of the frame is then scanned in $^1/_2 \times 520.000 = 260.000$ cycles. As the image is scanned 25 times per second, the frequency of the video signal is 25×260.000 c/s $= 6.5$ Mc/s A closer study shows that some correction is necessary, in consequence of which the highest modulation frequency used in practice need not exceed 5 Mc/s.

A number of European countries have accepted the 625-line system of the standard set up by the Comité Consultatif International des Radiocommunications (C.C.I.R.) [1]). According to this standard the video transmitter should be operated with an asymmetrical side-band, so that the frequency range used shall not be unnecessarily wide, i.e. the transmission should consist of the carrier wave and upper side-band, the lower side-band being for the greater part suppressed. The over-all width of the channel (sound and picture) is then 7 Mc/s.

The ideal frequency curve (aerial current as a function of the frequency) of the complete video transmitter is depicted in *fig. 1* (curve *1*); a certain tolerance is necessary, however, and this is represented by the difference between curves *1* and *2*, which means that at a frequency of, say 5 Mc/s above the carrier frequency, a drop of 50% = 6 dB is permissible.

The extent to which the frequency curve of a video transmitter is determined by its constituent parts depends on the method of modulation. The modulation systems in use may be grouped into two main classes, viz:

[1]) Standards for the international 625-line black and white television system, C.C.I.R., Geneva, 10th October 1950.

1) The R.F. drive voltage at the input of the output stage being kept constant, the grid bias is varied in accordance with the video signal; this is called output stage modulation.

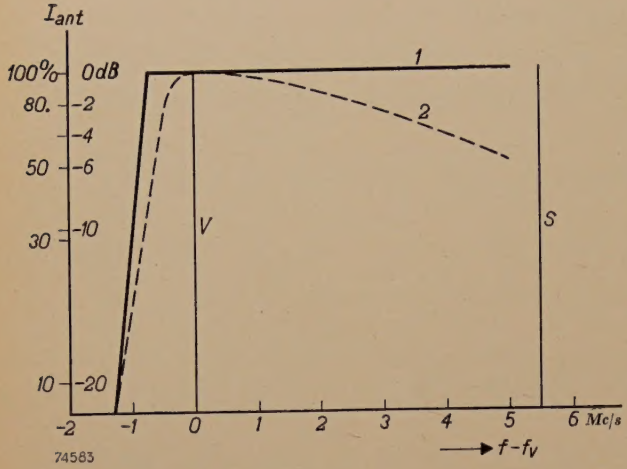

Fig. 1. *1*) Ideal, *2*) acceptable frequency characteristic of a television transmitter for 625 lines, in accordance with C.C.I.R. standard; aerial current I_{ant} (vertical); difference between the frequency f and the fixed frequency f_v of the video carrier wave (horizontal).
The vertical lines V and S represent the video and sound carriers respectively.

2) Modulation in one of the preceding stages, e.g. variation of the grid bias with the video signal in the control stage or one of the stages preceding it: the output stage then receives a modulated R.F. drive voltage and constant bias, and thus operates as a linear amplifier.

Now, the frequency curve of a transmitter with output modulation depends on the modulator, the driver and output stages, the aerial and the filter for suppressing the lower side-band (located between output stage and aerial). With modulation applied to one of the preceding stages, those stages which operate as amplifiers also affect the result.

In view of the object outlined above, viz. to generate considerable output power, this article is mainly concerned with the output stage of the transmitter. This part of the transmitter must be responsible for not more than a mere fraction of the permissible difference between the two curves shown in fig. 1. We have therefore set out to design an output stage tuneable from 48 to 68 Mc/s that will yield a resonance curve which does not drop more than 1 dB at a frequency of 5 Mc/s above the carrier frequency. This output stage is to consist of a single circuit, with resistive load. With a frequency range of 6 Mc/s (5Mc/s for the upper side-band plus 1 Mc/s for the lower), the current flowing in the loading resistor must therefore remain constant to within 1 dB.

That there is to be only one *LC* circuit means that the conditions are less favourable than in conventional transmitters, in which a transmission line, inductively coupled to the output stage, is used for feeding the aerial.

In order that the transmitter shall radiate as few harmonics as possible (as these may interfere with other channels), it is usual to couple this line inductively to the output circuit; the output and aerial circuits then function as a band-filter which, for the same valve loading, will give a greater resonance width than a single circuit.

This requirement can be met most simply by offsetting the tuning so that the resonance frequency does not coincide with the carrier frequency, but lies slightly higher, e.g. 1.5 Mc/s, as shown for the No. 4 TV channel [2]) in *fig. 2*. All the same, it is anything but simple to obtain a sufficiently flat characteristic over a range of 6 Mc/s, especially when high power is required. The conflicting conditions of a flat resonance curve and high power will now be examined.

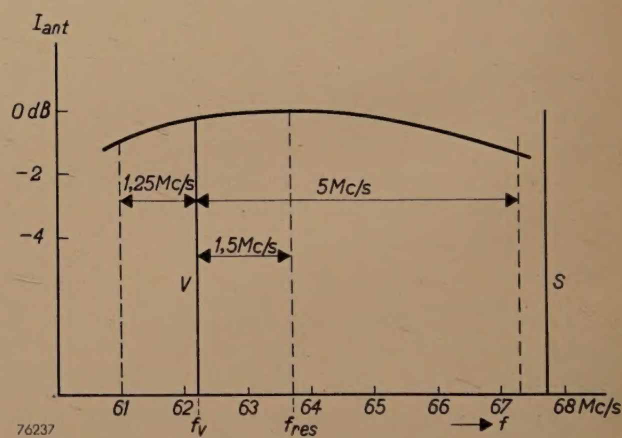

Fig. 2. Resonance curve of the output stage of a video transmitter. f_{res} resonance frequency, f_v frequency of the video carrier wave (TV channel 4). V video carrier, S sound carrier. A satisfactory region of the resonance curve with respect to the frequency band f_v — 1.25 Mc/s to f_v + 5 Mc/s is ensured by taking f_{res} to be roughly 1.5 Mc/s higher than f_v.

Resonance width and power

Fig. 3a shows the anode circuit of an output stage. For convenience the output circuit is represented by an *LC* circuit, although in television

[2]) At an international conference at Stockholm (1952) five plans were prepared for the distribution of wavelengths for European television transmitters in those countries which have accepted the international 625-line standard, and in Great Britain, France, Belgium and Eastern Germany. Some of the channels under the plan for the C.C.I.R. countries are mentioned in this article, hereinafter referred to as channels 2, 3 and 4 (the numbering is unofficial). The video carrier frequency of channel 2 is 48.25 Mc/s, that of channel 3 is 55.25 Mc/s and that of channel 4, 62.25 Mc/s; these all fall within the frequency band of 41 to 68 Mc/s mentioned above.

transmitters working at wavelengths of a few metres, a transmission line is usually employed (as mentioned later). The parallel resistance R represents the effective load of the aerial, as well as the losses in the circuit and the damping produced by the transmitting valves.

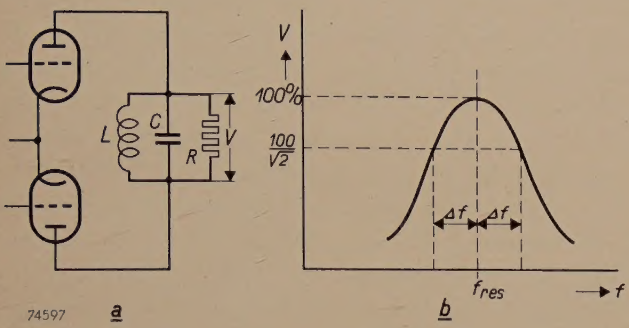

Fig. 3 *a*) Anode circuit of a push-pull output stage. *LC* oscillatory circuit with total load *R*.
b) Resonance curve: the voltage *V* across the network *LCR* as a function of the frequency *f* with constant alternating grid voltage. The resonance width is $2\Delta f$.

A circuit of this kind gives a resonance curve such as that shown in fig. 3*b*, the "width" of which is defined as $2\Delta f$, where Δf is the amount by which the frequency must differ from the resonance frequency for the voltage across the circuit to drop to $\frac{1}{2}\sqrt{2}$ times the peak value, the grid voltage remaining constant. For Δf we may write:

$$\Delta f = \frac{1}{4\pi CR}. \quad \ldots \ldots \quad (1)$$

Under normal operating conditions (R.F. Class B) the anode current is sinusoidal during one half-cycle (amplitude $I_{a\,\max}$), and zero during the second half (*fig. 4*). The amplitude of the first harmonic of this current (broken line) is $\frac{1}{2} I_{a\,\max}$ and the r.m.s. value is therefore $I_{a\,\max}/(2\sqrt{2})$. Hence, the power dissipated by the parallel resistance R is:

$$P = \left(\frac{I_{a\,\max}}{2\sqrt{2}}\right)^2 R,$$

or:

$$R = \frac{8P}{I_{a\,\max}^{\;2}}.$$

Substitution in equation (1) then gives for the half-width of the resonance curve:

$$\Delta f = \frac{I_{a\,\max}^{\;2}}{32\pi CP}. \quad \ldots \ldots \quad (2)$$

From this it is seen that Δf and P are inversely proportional. In order to secure the highest possible power for a given band-width, the capacitance C must be as small as possible and the peak current $I_{a\,\max}$ as high as possible. This peak current is limited by the saturation current of the cathodes. For the object in view, that is high power with large band-width, the valves used should therefore have a high saturation current, and this is obtained without entailing excessively high heater power by using thoriated tungsten cathodes; as explained in a recent issue of this Review [3]), this material can now also be used for high-power valves, owing to the introduction of more effective getters.

In the design of output stages the question arises whether triodes or tetrodes should be used, and one of the points to be considered in this choice is the effect of the internal resistance R_i. Let us take two examples, on referring to tetrodes and the other to triodes, for the same maximum anode dissipation. The output capacitance of these valves is roughly the same, and determines to a large extent the value of the capacitance C in equation (1). The resistance R will also be assumed to be the same for both types so that, in accordance with (1), the two circuits will have similar resonance widths. The resistance R may be regarded as consisting of three resistances in parallel, viz. resistance R_a representing the effective load of the aerial, $R_{L,C}$ corresponding to the losses in the oscillatory circuit, and another resistance to represent the damping due to the valves; under Class B conditions the last-mentioned is $2R_i$.In tetrodes R_i is so high that its effect can be disregarded. In triodes, however, the value of R_i is such that R_a must be appreciably higher than in tetrodes in order to sustain the assumption of equal total resistances R (the loss resistances $R_{L,C}$ are assumed equal). The higher value of R_a is an advantage, for the required bandwidth was obtained by choosing R_a lower than the value that matches the valve impedance; with the higher value of R_a now required the matching is improved, i.e. the output is increased.

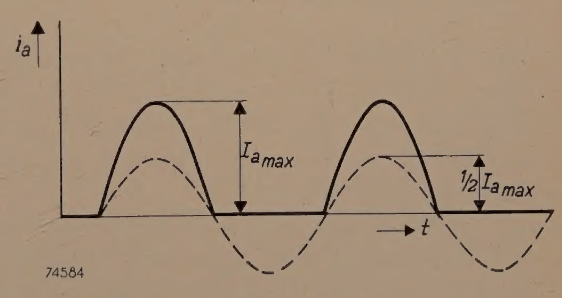

Fig. 4. Full line: anode current i_a of one of the valves in an R.F. Class B push-pull circuit, as a function of the time t. Dotted line: the first harmonic of i_a.

This advantage of the triode is offset by two drawbacks; firstly, the value of R_i varies in consequence of the modulation; this is referred to again at the end of the article. Secondly, triodes in general require more drive power than tetrodes of the same rating.

In practice both kinds of valve give satisfaction; tetrodes are used at the transmitter at Lopik (Netherlands). The tests described in the following all relate to triodes, however.

[3]) E. G. Dorgelo, Philips tech. Rev. 14, 226-234, 1953 (No. 8).

In order to ascertain whether a resonance curve such as that shown in fig. 2 can be obtained simultaneously with high output power, and also whether the modulation characteristic will then be satisfactory, we have constructed an experimental unit comprising an output stage with driver stage and preceding circuits, feeding a load resistance (dummy aerial). The frequency is variable from 48 to 68 Mc/s, which range covers television channels 2, 3 and 4.

We shall first say something about the output circuit, then pass on to the different sections and finally describe the results obtained.

Circuit of the output stage

An advantage of push-pull output circuits is that a point of symmetry occurs at the input of the circuit, at which there is no R.F. voltage with respect to earth. When modulation is applied to the output stage it can be introduced directly at this point, thus keeping the load on the modulator at a minimum. On the other hand, asymmetrical circuits (a single valve, or several in parallel) entail, in the range of frequencies with which we are concerned, the use of by-passing capacitances (coaxial systems), which constitute an extra load.

At the same time, it should be added that this capacitance can be avoided very simply by increasing the length of the coaxial system by one quarter-wavelength. However, this is practicable only when the wavelength is short enough (frequencies above 100 Mc/s); at such frequencies an asymmetrical circuit is therefore often employed.

It is necessary next to consider whether the valves are to be operated in an earthed-cathode, or earthed-grid circuit, i.e. whether the cathodes or the grids are to form the common point in the input and output circuits.

A conventional circuit for short-wave transmitters with small band-width is the earthed-cathode circuit shown in *fig. 5a*. To avoid coupling between the anode and grid circuit via the anode-to-grid capacitance, such that the stage oscillates, the grid of each valve is connected to the anode of the other through a neutralising capacitor C_n. As long as the frequency is comparatively low, coupling is completely eliminated when $C_n = C_{ag}$.

In an earthed-cathode circuit neutralised in this way there are two capacitive branches in parallel with the coil in the anode circuit, each consisting of C_{ag} and C_n in series. Since $C_n = C_{ag}$, these branches function as a single capacitance (C_{ag}). In the earthed-grid circuit, on the other hand (fig. 5*b*), we are mainly concerned with a branch in parallel with the anode coil, comprising the two capacitances C_{ag} in series, which behave as a capacitance $^1/_2\, C_{ag}$ (In both cases the slight effect of the capacitance C_{af} between anode and cathode may be ignored; C_{af} is roughly $^1/_{25}\, C_{ag}$.) With the earthed-grid circuit the contribution of the valves to the over-all output capacitance is thus roughly half that of the earthed-cathode circuit. This is of considerable importance where large transmitting valves are employed, as $^1/_2\, C_{ag}$ (or C_{ag}) then constitutes the greater part of the total output capacitance. A low capacitance is an advantage firstly from the point of view of the resonance width and output power rating (see Eq. (2)) and secondly, because the alternating current in the *LC* circuit is lower and the losses smaller for given frequency and amplitude of the R.F. voltage.

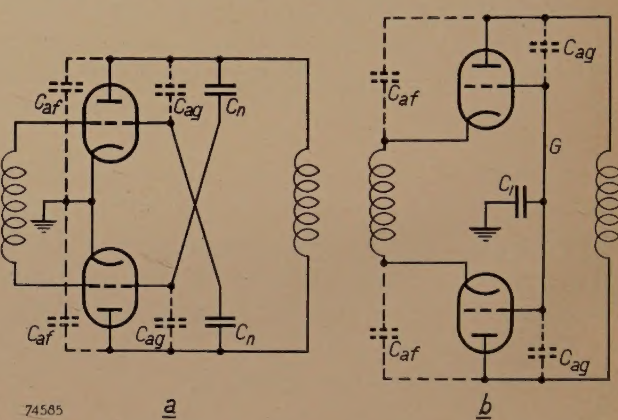

Fig. 5. Push-pull stage with *a*) earthed cathodes, *b*) earthed grids. C_{ag} anode-grid capacitance, C_{af} anode-cathode capacitance, C_n neutralising capacitor, C_1 capacitance between earth and a screen plate *G* used for connecting the grids.

In the neutralised, earthed-cathode circuit (fig. 5*a*) the self-inductance of the leads to the neutralising capacitor, at very high frequencies, affects the value that C_n should assume. The precise value of C_n thus becomes dependent on frequency to such an extent that insufficient neutralising is obtained throughout the frequency band of a video transmitter. This applies equally to any other neutrodyne circuit. In this respect, too, the earthed-grid circuit is much better than the earthed-cathode arrangement, provided that the valves are such that the grid serves as an effective screen between the input and output circuits; such screening should be not only electrical, but also magnetic. At higher frequencies the current flows only in the thin outer layer of the conductor (skin effect), and this effect can be put to good use in that, if matters are carefully arranged, the input and output circuits can be so separated that neutralising capacitors are no longer necessary.

Fig. 6 illustrates a suitable arrangement. The ring-seal grid contacts of the valves are mounted in a common metal screen plate. Owing to the skin effect, the

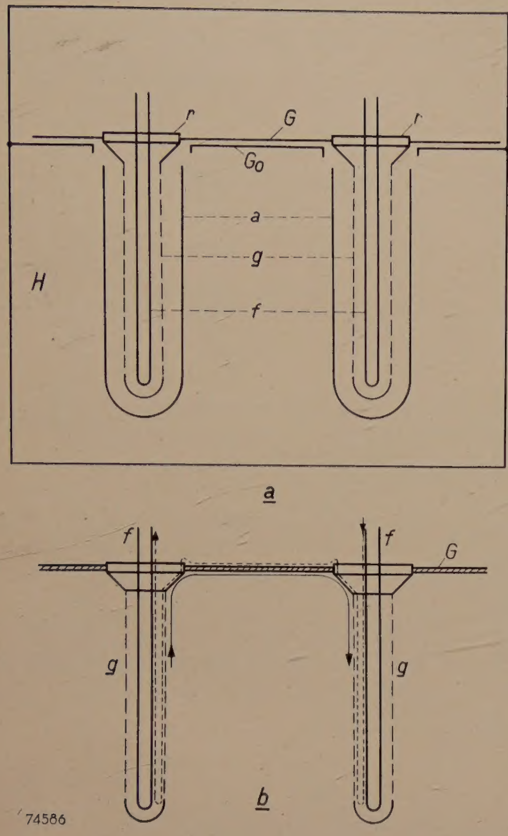

Fig. 6. *a*) Cross-sectional diagram of two transmitting valves with earthed grids. *a* anodes, *g* grids, *f* filaments, *r* ring-shaped grid seals connected to common screen plate *G*. The plate G_0, which is part of the chassis *H*, forms with *G* a capacitor (C_1 in fig. 5*b*).
b) Path of the R.F. currents through filaments, screen plate and grids. The input current flows in the skin shown by the dotted arrow, and the output current in the skin indicated by the full arrow. The only residual coupling between the circuits is that which occurs by way of the mesh of the grids.

input current — as far as the grids and screening plate are concerned — flows only on the upper face of the plate and the inner faces of the grids, and the output current only on the underside of the plate and the outer faces of the grids. The two currents are thus separated by the body of the screen plate and grids.

Imperfections in the screening occur only as a result of the presence of the grid meshes; these give rise to a certain capacitance C_{af} between anode and cathode, which is accompanied by mutual inductance between the circuits. Below a certain frequency the coupling across C_{af} is dominant; above that frequency, the coupling through the mutual inductance becomes the more important. This is shown by the so-called "interaction" test in which an RF voltage of variable frequency and constant amplitude is applied between the cold cathodes. The RF voltage in the output circuit resulting from the imperfect screening is then measured at the output as a function of the frequency. The result of such a test, as applied to the output circuit of our experimental equipment arranged as shown in fig. 6, is demonstrated in *fig* 7. A minimum occurs at 62 Mc/s; at that point the effect of C_{af} counterbalances that of the mutual inductance. (The minimum is not at zero as a result of losses.) Throughout the whole frequency range from 48 to 68 Mc/s (channels 2, 3 and 4), this interaction is so slight that the stage under discussion is stable in operation, without taking further measures to ensure this.

As the grids have to be biased, they cannot be connected directly to the chassis, but only via a capacitor. As will be seen from fig. 6, this capacitor is formed by the screening plate and a second metal plate which is actually part of the chassis itself; these plates are separated by a solid dielectric such as mica or polythene.

Although our experimental unit is not adapted for modulation, a few remarks on this subject will not be out of place.

The arrangement shown in fig. 6 is suitable for use as an amplifier, the modulation being therefore applied to a preceding stage; the modulated R.F. drive voltage is applied between the cathodes of the output valves. For modulation in the output stage itself, the video signal, in this circuit, is applied between the cathodes and earth, but this has the disadvantage that the modulator is very heavily loaded, not only

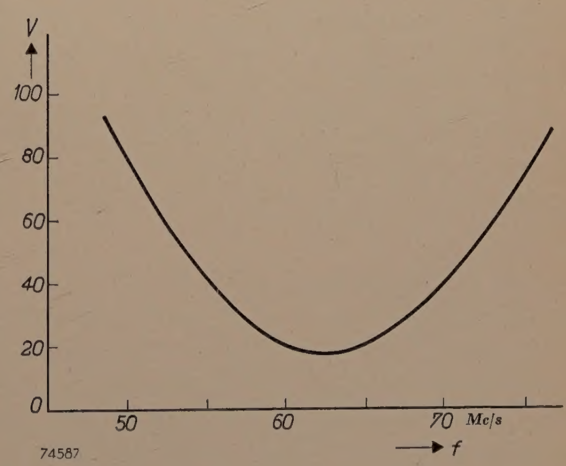

Fig. 7. Interaction: measured voltage *V* between the anodes of the output stage (on relative scale) as a function of the frequency *f*, with constant input voltage and cold cathodes.

by the valve capacitances, but also by the capacitance of the filament transformer. Preference is therefore sometimes given to the circuit shown in *fig. 8a* in which the grids are not connected directly to each other, but with two capacitors in series between them (for symmetry), and a choke with a centre tap to which the video signal is applied. Fig. 8*b* shows a practical form of the circuit.

Owing to the absence of a continuous screening plate, the paths of the current in the input and output circuits are not fully separated, and a certain amount of coupling occurs. This can be represented, together with the magnetic coupling due to the meshes of the grid, by a self-inductance L' in series with each grid (fig. 8*a*). The equivalent network for one half of the circuit is depicted in fig. 8*c*, in which the valve capacitances (C_{ag}, C_{gf} and C_{af}) are also shown. It can be shown that the coupling between the input and output circuits is zero when

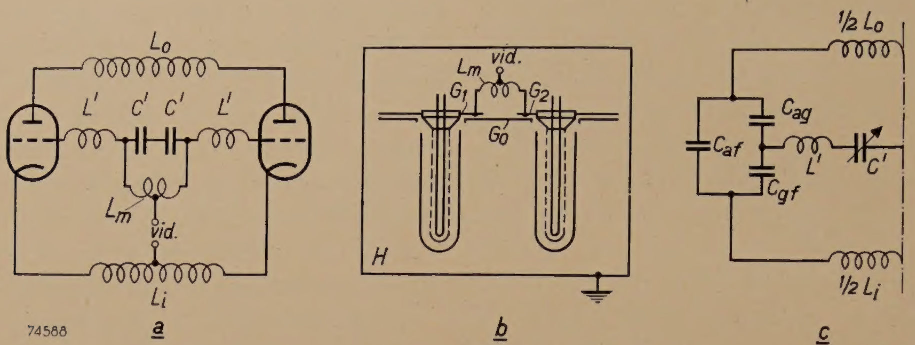

Fig. 8. a) Circuit for modulation in the output stage. The grids of the valves are connected to each other capacitively (C'-C') and the video signal *vid* is applied through a centre-tapped choke (L_m). L_i input circuit, L_o output circuit. L' stray inductance.
b) Practical form of circuit *a*). The capacitors C' are formed by screening plates G_1 and G_2 on the one hand and plate G_0 (not connected to the chassis H) on the other.
c) Equivalent circuit of one half of the arrangement. C_{ag}, C_{gf} and C_{af} valve capacitances; L' stray inductance. Coupling between circuits can be reduced to zero by a suitable value of C'.

$$X' = \frac{-X_{fg}\, X_{ag}}{X_{fg} + X_{ag} + X_{af}},$$

where $X' = \omega L' - \frac{1}{\omega C'}$, $X_{fg} = \frac{1}{\omega C_{fg}}$, etc. (with $\omega = 2\pi f$).

This equation is satisfied by giving C' a certain value. This "neutrodyning" is again dependent on frequency, to such a degree that stabilization is achieved only in a small range of frequencies. In Great Britain, where the 405-line system is employed and the frequency band is accordingly fairly narrow, the latter method has been adopted, but in transmitters operating with 625 lines or more it is less suitable owing to the high "selectivity" of the neutrodyne.

When comparing the merits of earthed-cathode and earthed-grid circuits we must also take into consideration the driver stage. The driver stage of an earthed-cathode circuit delivers power which serves only to cover the grid dissipation in the output valves and grid resistors, together with the losses in the input circuit. On the other hand, in the earthed-grid circuit the driver stage must deliver a certain amount of power which is passed on to the load in the output stage. This is illustrated in *fig. 9*, which shows the simplest form of earthed-grid circuit. The A.C. voltage across the load resistance R is the sum of the alternating input voltage V_i and the alternating voltage across the valve V_a. If we denote the anode alternating current by I_a and if I_a, V_i and V_a be regarded as r.m.s. values, the power delivered is $I_a(V_i + V_a)$; of this, apart from the losses already mentioned, the quantity I_aV_i must be supplied by the driver stage, which thus has to be larger than for an earthed-cathode output stage. This disadvantage, however, does not outweigh the above-mentioned advantage of the lower output capacitance and greater stability of the earthed-grid circuit.

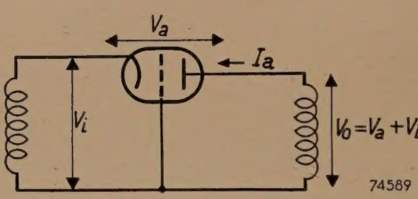

Fig. 9. Amplification stage with earthed grid. The alternating output voltage V_0 is the sum of the alternating input voltage V_i and the alternating voltage V_a between anode and cathode; the power applied to the input is thus augmented at the output (apart from losses).

Transmission lines

Oscillatory circuits for metre wavelengths are best made in the form of transmission lines, either as parallel line or coaxial conductors. Both types are employed in the present equipment. The parallel lines may be seen in the diagram of the output stage in *fig. 10*, in which G is the screening plate mentioned above, to which the grids are connected. The cathodes are connected to one end of a two-wire line (Le_i) to form the input, the other end being shorted by a bridge piece B_1. Because of the relatively high input capacitance of the valves, the filaments themselves constitute a not unimportant part of the input circuit, and the external conductors must therefore be very short. The driver stage (S) is connected to a suitable point on these

conductors, which has to be sufficiently far removed from the shorting piece. To prevent the tapping point from falling too near the filament connections (and thus being inaccessible), the input circuit is extended by one half-wavelength, to about $^3/_4$ wavelength. Tuning is effected by moving the shorting bridge.

self-inductance [4]), and adjustment of the shorting bridge varies the self-inductance and therefore also the natural frequency of the circuit. As we have already shown above, the effective capacitance in the circuit is, in the main, $^1/_2\,C_{ag}$, since the two anode-to-grid capacitances are in series. (Strictly speaking, C_{ag} is not concentrated at any one point,

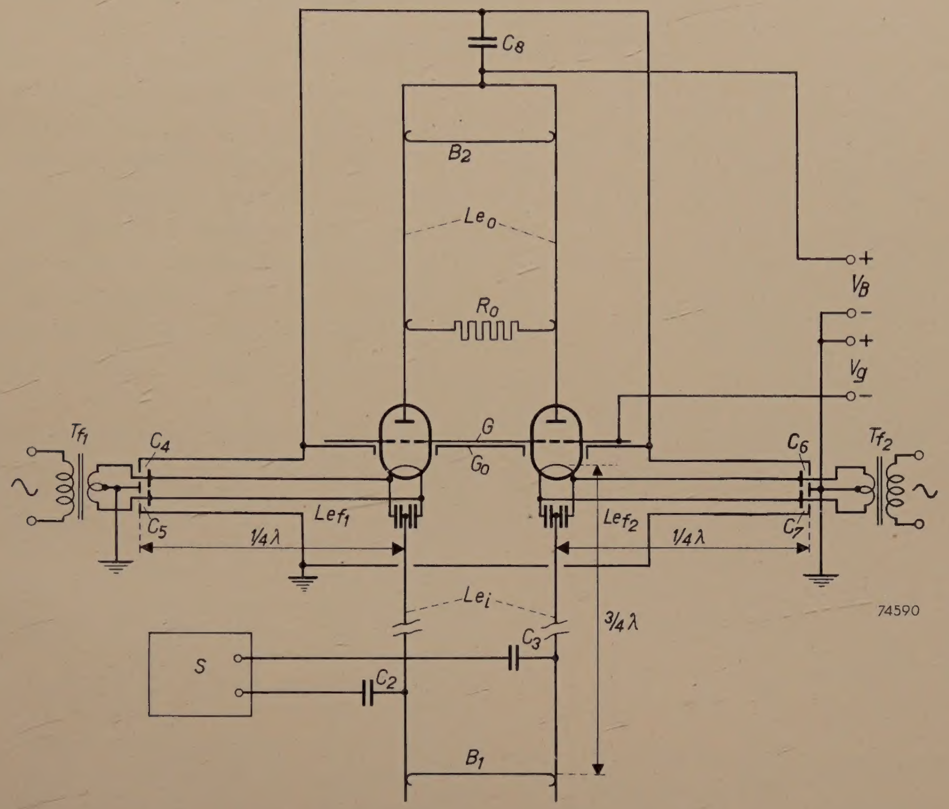

Fig. 10. Circuit diagram of the output stage. S driver stage coupled through capacitors C_2 and C_3 to the input transmission line Le_i with tuning bridge B_1. Lef_1, Lef_2 transmission lines through which transformers Tf_1 and Tf_2 feed the filaments. C_4-C_5 and C_6-C_7 capacitive tuning bridges for these transmission lines. G screen plate. G_0 earthed plate (Cf fig. 6). Le_o output transmission line with tuner B_2 and movable loading resistor R_0. C_8 by-pass capacitor. V_g D.C. grid voltage. V_B H.T. supply.

The filament transformers (Tf_1,Tf_2) are connected by a quarter-wavelength coaxial cable (Lef_1, Lef_2). At one quarter-wavelength from the filament, the filament leads are connected capacitively to the outer conductor (capacitances C_4-C_5 and C_6-C_7). The secondary winding of the filament transformer is earthed.

The anode or output circuit constitutes a fourth transmission line, but in this case the arrangement is more complicated, since the valves are an integral part of the system. Each of the transmitting valves is contained in a cylinder (C, *fig. 11a*). The two cylinders together function as a short two-wire line, shorted by a movable bridge (B_2 in figs. 10 and 11a). As is known, a line of this kind behaves like a

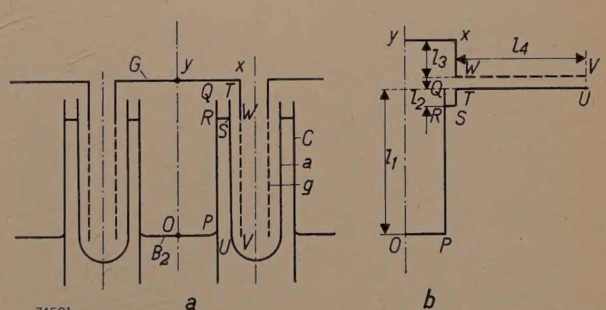

Fig. 11.a) Output transmission line. g grid, a anode, C metal cylinder enclosing anode, with movable tuning bridge B_2, G screen plate.
b) Arrangement of the two-wire and coaxial lines showing lengths l_1, l_2 and l_3, l_4. The points O, P, Q, ... X, Y correspond to those in (a).

4) For this and other properties of transmission lines see article by C. G. A. von Lindern and G. de Vries, Philips tech. Rev. 6, 240-249, 1941.

but is the equivalent value of the open coaxial line formed by the anode and grid). Besides this, there is the capacitance of the two-wire line represented by the two cylinders, but as this line is short compared with the wavelength λ (roughly $^1/_{15}\ \lambda$), the capacitance is negligibly small.

The transmission line formed by the anode circuit is shown diagrammatically in fig. 11*b*. Here are seen a shorted two-wire line (the two cylinders and the shorting plate) of which the length is l_1, two shorted coaxial lines (l_2 and l_3) and an open line, shown in the figure as a two-wire line, representing the grid and anode, of length l_4. According to equations (19) and (24) in the article referred to in footnote [4]) the input impedance Z_k of a shorted line — provided length l is small compared with the wavelength (as in this case) — may be expressed as:

$$Z_k = j\,\zeta \tan 2\pi \frac{l}{\lambda}$$

and the input impedance Z_0 of an open line as:

$$Z_0 = -j\,\zeta \cot 2\pi \frac{l}{\lambda},$$

where ζ is the characteristic impedance of the line, a real and positive quantity which is a known function of the dimensions. From these formulae it is seen that the shorted line does indeed function as a self-inductance, and the open system as a capacitance.

The natural frequencies of an electrical system are those at which current can flow in the absence of an electromotive force, that is, when the sum of all the reactances is zero. Applying this to the system depicted in fig. 11*b*, we find that:

$$j\,\zeta_1 \tan 2\pi \frac{l_1}{\lambda_{\rm res}} + j\,\zeta_2 \tan 2\pi \frac{l_2}{\lambda_{\rm res}} + j\,\zeta_3 \tan 2\pi \frac{l_3}{\lambda_{\rm res}} - -j\,\zeta_4 \cot 2\pi \frac{l_4}{\lambda_{\rm res}} = 0.$$

This expression can be employed for estimating the natural frequency $\lambda_{\rm res}$ for given dimensions.

The curve shown in *fig. 12* shows how the measured resonance frequency of the anode circuit depends on the position of the shorting bridge; it will be seen that the system can be tuned to TV channels 2, 3 and 4, the video carriers of which have frequencies of 48.25, 55.25 and 62.25 Mc/s respectively. As mentioned above (fig. 2), it is an advantage to choose the resonance frequency of the anode circuit somewhat higher.

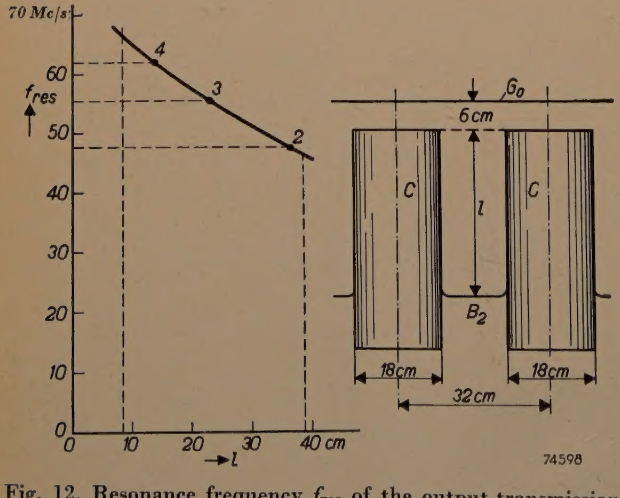

Fig. 12. Resonance frequency $f_{\rm res}$ of the output transmission line plotted against distance l of the tuner B_2 from the top edge of the cylinders C (see fig. on right). The points 2, 3 and 4 correspond to the carrier frequencies of TV channels 2, 3 and 4.

Constructional details

The output stage

Two type TBW 12/100 valves are used in the output stage. This valve was described fully in a recent issue of this Review (see article referred to in footnote [3]). The arrangement of the grid is such that the valve is suitable for earthed-grid circuits; the cathode is of thoriated tungsten and the anode is water-cooled. This valve was designed for operating at frequencies up to 30 Mc/s, but in our test it was used successfully at frequencies up to about 70 Mc/s.

Fig. 13 depicts the actual output stage and shows the two cylinders in the output circuit, with tuning bridge.

A cross-section of one of the cylinders, with valve in position, is shown in *fig. 14*. That part of the grid which lies above the anode seal is in effect a cylinder with no apertures in the wall; the anode is extended some distance above the seal, which prevents the glass from being struck by the electrons. Cooling air is directed towards the anode and grid-seal from a perforated ring R which is standard with the valve TBW 12/100. The coaxial cylinder is extended right up to this ring, which results in the following advantages:

1) Since the potential is the same at all parts of the cylinder at the same height, the anode seal is then in a field-free space and no dielectric losses or consequent heating occurs in the glass.
2) The greater length of the cylinders facilitates adjustment of the loading resistor (see below).
3) Proximity effect at the anode seal is reduced.

The last-mentioned point can be elucidated as follows. As is known, skin effect occurs at high frequencies, this being manifested as an increase in resistance; it is actually a result of non-uniform current distribution due to the magnetic field of the current itself. Proximity effect occurs between two or more conductors, because the distribution of the current in the one is influenced by the magnetic field of the other(s) [5]). The current density is thus

[5]) See for example Philips tech. Rev. 6, 220, 1941, or F. E. Terman, Radio Engineers' Handbook, McGraw Hill, New York, 1943, pp. 36-37.

increased at those sides of the conductors which face each other, and this results in a further increase in the resistance. When two valves are mounted close to each other without the screening effect of the cylinders already mentioned, the high-frequency anode current would not be uniformly distributed round the periphery of the anodes, but would be highly concentrated at the sides which face each other, and this would entail some risk of overheating at the anode seals. The above-mentioned extension of the cylinders ensures that the current distribution is much more uniform. Proximity effect does occur in the cylinders themselves, but this can do no harm.

Not only the tuning bridge, but also the load resistor (R_0 in figs. 10 and 13) can be moved along the cylinders. The voltage between points at equal height on the cylinders increases from the bottom upwards, so that the resistance forms a greater load according as it is pushed upwards; this provides a continuously variable load with constant resistance.

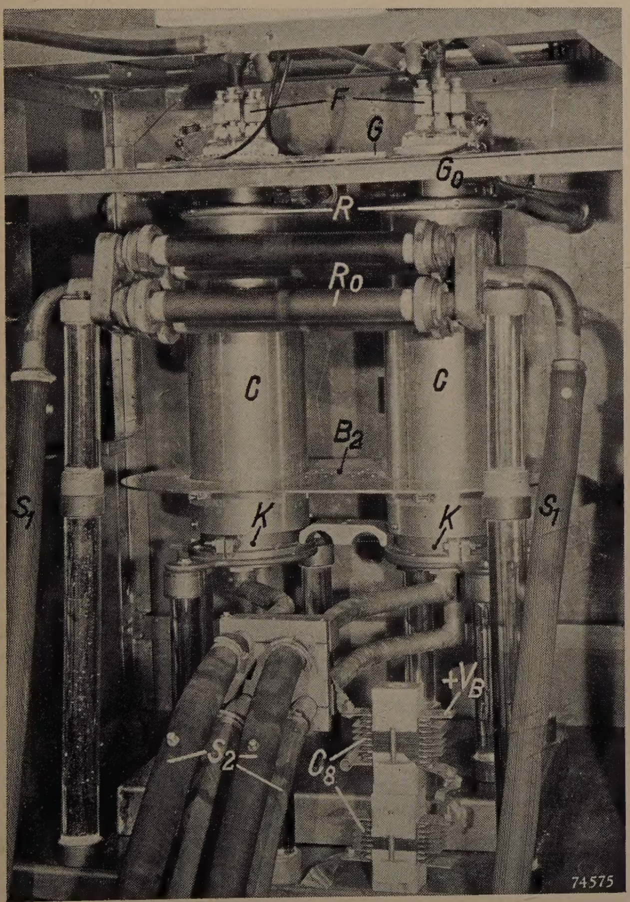

Fig. 13. Photograph of the output stage. F filament connections of the valves (TBW 12/100). G screen plate connecting the grids and insulated from plate G_0 which is in contact with the chassis (see fig. 10). R perforated rings through which cooling air is forced against the glass part of the transmitting valves. C cylinders of the output transmission system, with tuner B_2 and loading resistor R_0, both movable. K cooler; S_1, S_2 cooling-water hoses for the loading resistor and valves. $+V_B$ supply voltage terminal. C_8 by-pass capacitor.

The load resistor itself consists of four ceramic tubes carrying a layer of sintered carbon, and through which cooling-water is run. With a flow of water of 25 litres/min, 25 kW per resistor tube can be dissipated. The four resistors are connected in parallel and, as their length is about 0.1 λ, the resistance value (120 Ω) is practically independent of frequency.

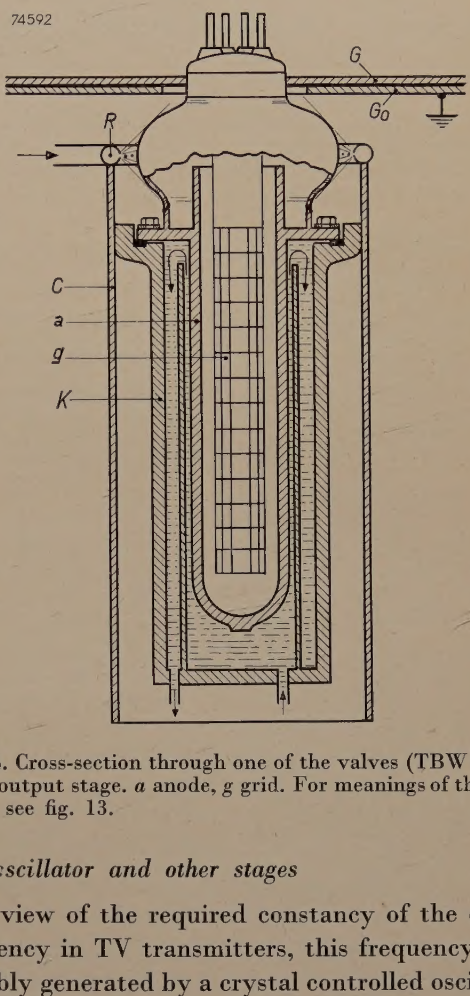

Fig. 14. Cross-section through one of the valves (TBW 12/100) in the output stage. a anode, g grid. For meanings of the other letters see fig. 13.

The oscillator and other stages

In view of the required constancy of the carrier frequency in TV transmitters, this frequency is invariably generated by a crystal controlled oscillator. Between the oscillator (which does not deliver very much power) and the output stage, a number of stages of R.F. amplification are required. In our experimental equipment, however, a variable frequency was needed, for which reason the stage for driving the output stage has been made up as a variable-frequency driver. The amount of power required for driving an output stage delivering 100 kW is 15 to 20 kW, and it is not so simple to construct an oscillator to give this amount of power, without parasitic oscillation, and giving a reasonably constant output voltage within the whole range from 48 to 68 Mc/s. The arrangement mentioned above, consisting of an oscillator of low power with subsequent stages of amplification, is by far the better of the two. This method was therefore

adopted, the frequency being of course variable; see block diagram in *fig. 15*.

The oscillator (O) has two type TB 2.5/300 valves [6]) in push-pull. To vary the frequency either the capacitance or the self-inductance can be varied. An LC circuit with variable C is incorporated

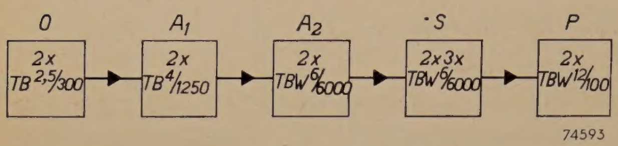

Fig. 15. Block diagram of the experimental equipment. O oscillator with variable frequency. A_1, A_2 amplifiers. S control stage, P output stage. The numbers of valves and the valve types are shown in the blocks.

between the grids, with another LC circuit, of which L is variable (variometer) between the anodes. The output voltage thus obtained is sufficiently constant throughout the whole range of frequencies. The oscillator is provided with a frequency-calibrated scale.

Between the oscillator and the driver stage two R.F. amplifiers or sub-driver stages are also necessary (A_1 and A_2, fig. 15). These are operated with two TB 4/1250 and two TBW 6/6000 valves in earthed-grid push-pull circuits.

The driver stage proper (S, fig. 15) is also in push-pull with earthed grids, using three type TBW 6/6000 valves on each side (*fig. 16*). The cooling-water inlet and outlet system consists of concentric tubes, the outer one of which simultaneously serves as the transmission line from which the output stage is driven. As in the output stage, use is made of a screening plate connected to the grids and earthed capacitively. The only reason why each valve has its own screening plate (as will be seen from fig. 16), is that it is thus possible to measure the grid current of each valve separately.

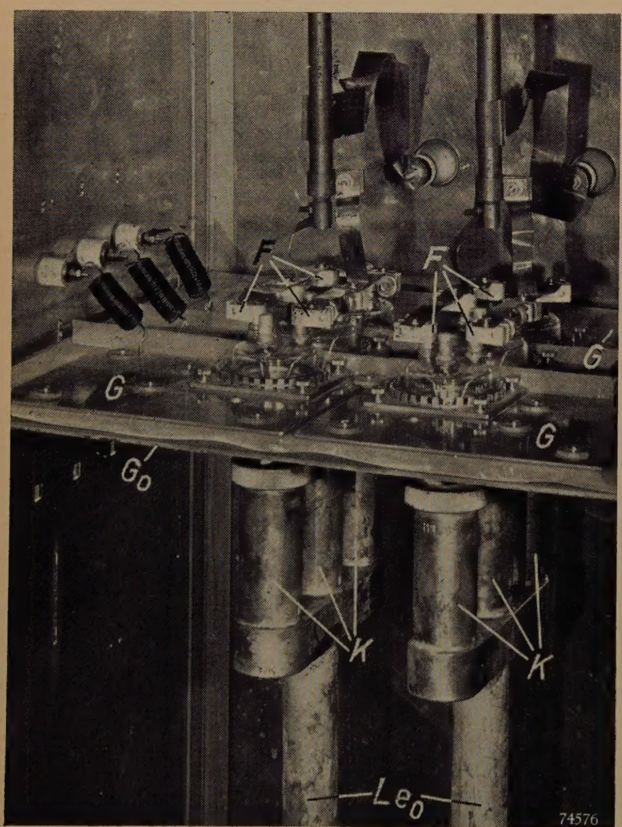

Fig. 16. Interior view of the driver stage with three type TBW 6/6000 valves on each side of an earthed-grid push-pull circuit. F filament connections. G screening plate. G_0 plate in contact with panel. K coolers. Le_0 output transmission line.

Measuring facilities

Measurement of the output power as a function of the frequency must be carried out with constant input voltage at the output stage. In effect, then, we measure the input voltage and the output power. As already mentioned, the frequency is read from the scale on the oscillator.

Measurement of the input voltage

The diagram in *fig. 17a* illustrates the method of measuring the input voltage. By reason of the capacitance between one of the input lines (Le_i) and a copper probe plate (D) mounted in the vicinity, RF voltage occurs on this plate. This voltage is passed to a germanium diode (Ge), a smoothing resistor(R) and a capacitor (C), and is measured on a relative scale by means of a D.C. instrument (microammeter).

The difficulty of a measurement of this kind is one of avoiding the interference from the strong magnetic and electric alternating fields in the region of the output stage, and the method employed to prevent such interference will be seen in fig.17*b*, which shows a diagram of the measuring equipment. The probe plate is located in or near the open end of a screening tube which contains the germanium diode, the resistor and the capacitor; the plate can be moved away from, or drawn into the tube, to suit the required sensitivity. A flexible coaxial cable connects this unit to the microammeter, which is placed in a metal box to screen it from R.F. currents which might otherwise damage it.

Another point for consideration is the frequency-dependence of the measurements. When the fre-

[6]) A description of the transmitting triode TB 2.5/300 is given in Philips tech. Rev. **10**, 273-281, 1949.

[7]) The TB 4/1250 is a valve similar to the TB 2.5/300, but has a higher power rating; the TBW 6/6000 is similar to the TBW 12/100, but is of lower power rating.

quency is varied the potential maxima and minima travel along the transmission line and the deflection of the meter changes. Now, the voltage in the direction of the transmission line varies considerably in the region of a minimum, but only slightly in the vicinity of a maximum, and it is therefore an advantage to mount the probe plate at a point where a maximum occurs. If this is done, the error due to displacement of the maximum, within the frequency range of a single channel, can be disregarded. For other channels the position of the relevant maximum has first to be located.

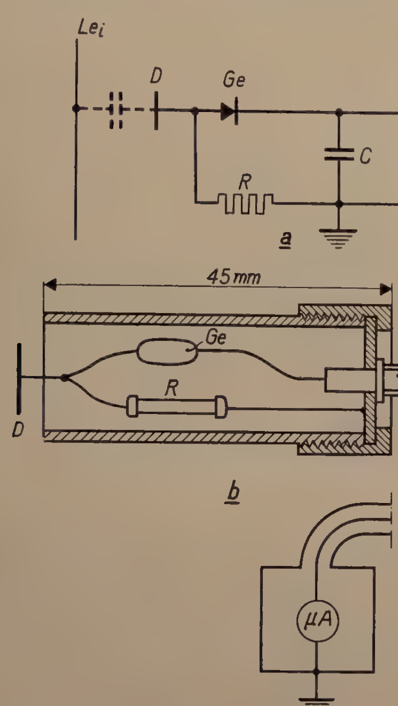

Fig. 17. *a*) Measuring circuit for the alternating input voltage V_i. Le_i one of the conductors of the input transmission line, D probe plate. *Ge* germanium diode. R and C smoothing resistor and capacitor. μA microammeter.
b) The assembled unit. Components D, *Ge*, R and C are housed in an earthed metal tube for screening purposes. The meter is connected by means of coaxial cable and is also screened.

As pointed out, the measurement is only relative, but this is sufficient if its sole purpose is to ensure that the input voltage remains constant. With a little extra trouble the equipment could be calibrated so that the microammeter gives direct indication of the R.F. voltage between cathode and grid. In so doing we would be making use of the fact that grid current commences to flow when the peak R.F. voltage between cathode and grid is practically equal to the D.C. voltage between these electrodes. This D.C. voltage (V_g in fig. 10) is variable and can be directly measured. The occurrence of grid current can be observed by means of a D.C. instrument in series with the voltage source V_g.

Measuring the output power

With constant water flow the increase in temperature of the cooling-water in the load resistor R_0 is a measure of the power P_0 dissipated. The temperature of the inlet water being constant, the thermometer used for measuring the temperature at the water outlet may be calibrated in kW to give a direct reading of P_0.

Results obtained with the experimental equipment

It is found that the equipment is capable of delivering a continuous output of 100 kW. The suitability of such equipment, when complete with modulator, aerial, etc., as a television transmitter for 100 kW peak output, can be confirmed only when it is known that the bandwidth is sufficient and the modulation curve sufficiently linear. Results concerning these details are given below.

As a basis we have taken the C.C.I.R. standard, case which specifies negative modulation [8]). In this the aerial current is at its maximum (100%) for the peak of the synchronisation signals; it drops to 75% on black level (lowest image brightness) and to 10% on white level (highest image brightness). Hence, at black level, the transmission power is $0.75^2 = 56\%$ of the peak power. As the TBW 12/100 valves can deliver 100 kW continuously, there is therefore a very wide margin of safety with 100 kW at the sychronisation peaks.

Resonance curves

Measurements were taken of the temperature increase $\Delta\vartheta$ of the cooling-water in the load resistor as a function of the frequency, in the region of the resonance frequency; this was effected with constant tuning of the anode circuit and constant input voltage, for different values of the load, which was varied by changing the position of the resistor R_0.

Figs. 18a and *b* show the "aerial current" $I_{\mathrm{ant}}(\sim\sqrt{\Delta\vartheta})$ plotted against the frequency for TV channels 2 and 4. At resonance, I_{ant} was found to be 75% of the value relevant to 100 kW; hence these figures correspond to black level in a transmitter with 100 kW peak output. It appears that within a frequency range of 6 Mc/s, I_{ant} is less than 1 dB below the resonance value, thus meeting the requirement originally laid down.

As pointed out in the foregoing, this requirement is rather on the high side, but we consider that it should be, for the following reason. With modulation, the anode direct current is

[8]) See article mentioned in footnote [1]), or Philips tech. Rev. **13**, 313, 1952 (No. 11).

lower than the value relating to black level, and this is accompanied by an increase in the internal resistance R_i of the valves. The valve damping thus drops and the bandwidth is accordingly reduced. Hence, when the modulation depth is increased, the bandwidth decreases.

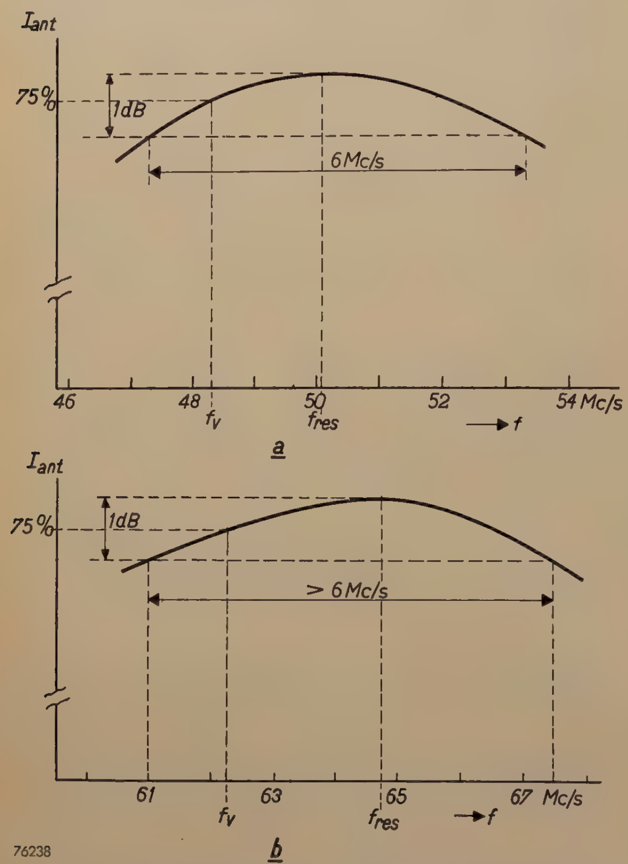

Fig. 18. Resonance curve I_{ant} = f (f) of the output stage tuned to *a*) TV channel 2, *b*) channel 4. Within a frequency range of 6 Mc/s, I_{ant} is less than 1 dB below the maximum value.

Modulation curves

In *fig. 19* the curves of various currents plotted against the alternating input voltage V_i are reproduced, viz. the "aerial current" I_{ant}, the anode (direct) current I_a, and the grid current I_g. Where I_{ant} = 100%, P_0 = 100 kW. The curve for I_{ant} is the modulation characteristic for modulation applied to one of the preceding stages (modulated input voltage); the analogous curves for modulation of the output stage (V_i constant, V_g variable) are shown in *fig. 20*.

To ensure satisfactory contrast (gamma) of the image as received, the curve for I_{ant} between black and white levels (75% and 10%) should be reasonably linear in both cases. Although nothing has yet been officially laid down as regards this linearity, it may be assumed that the curves as shown would be sufficiently linear. (It is to be expected that ultimate requirements will not be so stringent as

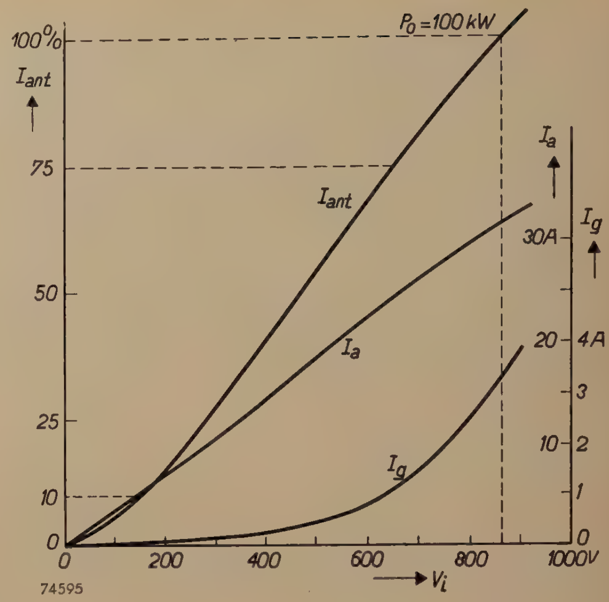

Fig. 19. Output curves for modulation in a preceding stage. "Aerial current" I_{ant}, anode (direct) current I_a and grid (direct) current I_g (I_a and I_g for the two TBW 12/100 valves together) plotted against the R.F. input voltage V_i (maximum peak between cathode and grid). D.C. grid voltage V_g = − 250 V. Supply voltage V_B = 6500 V. Loading resistance R_a = 350 Ω. Frequency f = 48.25 Mc/s (T.V. channel 2). I_{ant} = 100% corresponds to synchronisation peaks P_0 = 100 kW), I_{ant} = 75% (Cf. fig. 18) to black level and I_{ant} = 10% to white level.

those covering sound transmission, since the eye is very much less sensitive to non-linearity than the ear.) If necessary, the modulator can be designed with non-linear characteristics such that the non-linearity in the modulation curve is largely compensated.

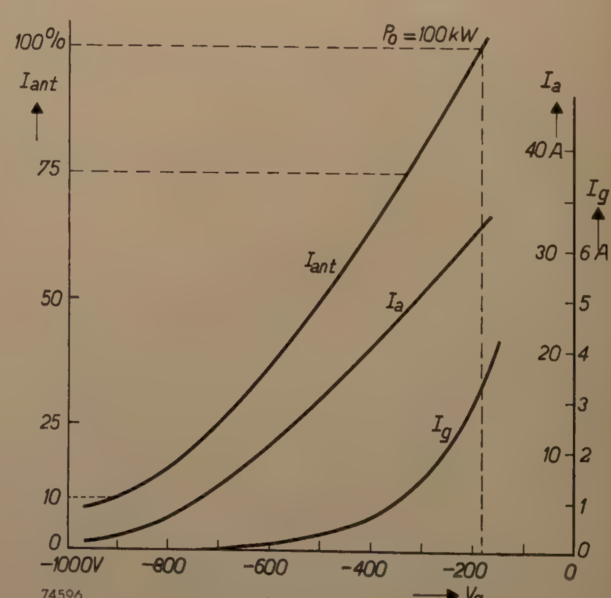

Fig. 20. Output curves for modulation in the output stage. I_{ant}, I_a and I_g as function of V_g, with V_i = 770 V, V_B = 6500 V, R_a = 350 Ω, f = 48.25 Mc/s. I_{ant} = 100% again corresponds to P_0 = 100 kW.

From the anode current curves in figs 19 and 20 it is seen that at $P_0 = 100$ kW the anode takes a current of 32 A, which, at 6500 V D.C., represents 210 kW. The grid current curves show that the control stage has to deliver current peaks of considerable amperage.

Summary. An experiment has been carried out to see whether a television transmitter can be designed for high power and having adequate bandwidth for a system of 625 lines. The object was to obtain, by means of an output stage consisting of a single circuit, a resonance characteristic which would be flat within 1 dB throughout a range of 6 Mc/s, using equipment intended to deliver 100 kW peak.

A description is given of the experimental equipment, in which the frequency is variable between 48 and 68 Mc/s (approx.); this range covers three television channels of 7 Mc/s band-width each. The equipment is capable of delivering 100 kW continuously, this being regarded as the peak value of the power that would occur during transmission of the synchronisation signals, with negative modulation. Resonance curves were plotted in respect of the corresponding black level, and these were found to meet the requirements imposed. Modulation curves were also plotted, for modulation in the output stage and in preceding stages, and these curves are reasonably linear between black and white levels.

The installation consists of an output stage for 100 kW and includes two TBW 12/100 valves in a push-pull earthed-grid circuit, preceded by a 20 kW driver stage with six type TBW 6/6000 valves, 2 sub-driver stages, and a low-power oscillator with variable frequency. The output stage was loaded with a water-cooled resistor, and the value of the power dissipated in this resistor was derived from the increase in temperature of the cooling-water.

A WEST-EUROPEAN TELEVISION NETWORK ON THE OCCASION OF THE CORONATION CELEBRATIONS IN LONDON

621.397.743

Fig. 1. Map of the provisional European television network. The circles represent television transmitters, the black dots radio-link transmitters and receivers.

The Coronation of Queen Elizabeth II on 2nd June 1953 was the occasion for the most ambitious television broadcast yet attempted in Europe. For this event a provisional international network was set up over which the British programme could be relayed to France, the Netherlands and West Germany.

The network itself (*fig. 1*) was made up of a chain of radio-links which relayed the vision signals to the various television transmitters. The sound signal was sent by land-line.

Originating in London, the signal was relayed in four steps to Cassel and thence in two further stages to Paris. Some of this part of the network had been already set up for previous transmissions between London and Paris. From Cassel a branch chain of radio-links took the signal to Breda in Holland, the final link with the Dutch television transmitters at Lopik and Eindhoven. From Eindhoven the signal was sent via a further link station (Helenaveen) to Germany. The radio-link equipment from Lille to Lopik and Helenaveen was provided by Philips Telecommunication Industries of Hilversum.

Fig. 2. Trailer housing the 405-625 "line-converter", at the foot of the church-tower in Breda on which a radio-link receiver and two transmitters are installed.

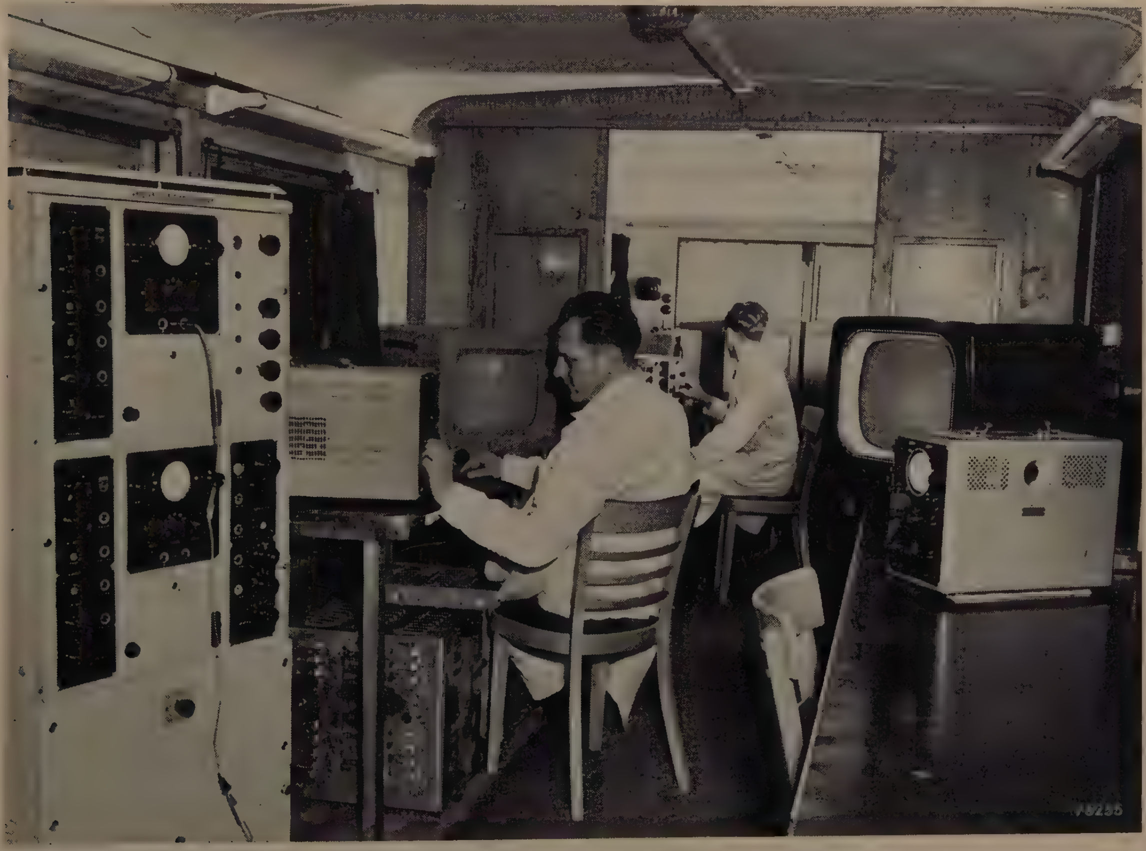

Fig. 3. Interior of the trailer stationed at Breda. The line-converter, not clearly visible in this photograph, is at the rear. Next to it are monitors for the 405 and the 625-line pictures. At the left in the foreground is the cabinet containing the generators of the line and frame frequencies for the 625-line system. The frame frequency was derived from the line frequency by a frequency divider, and synchronised with the frame frequency of the British system.

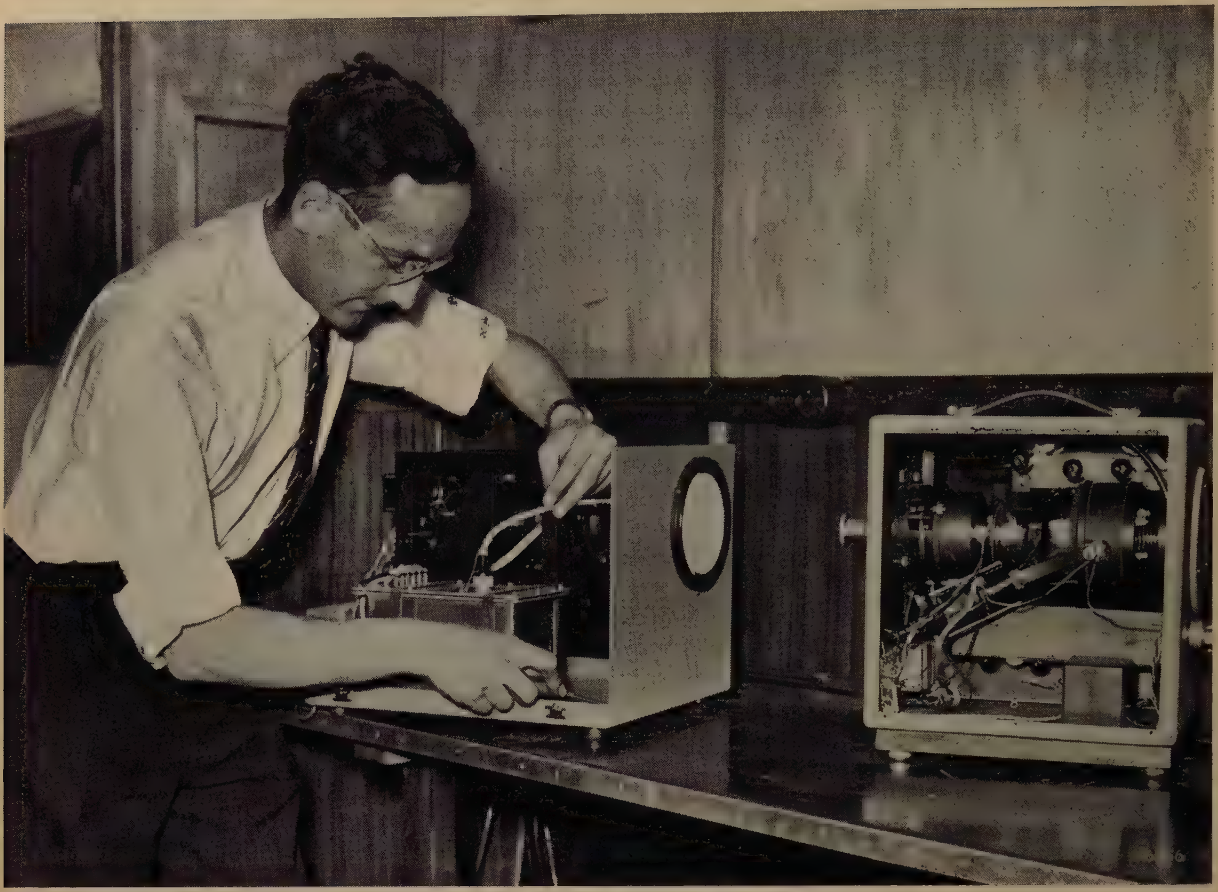

Fig. 4. The 405-625 line-converter used at Breda. On the left is the unit containing the picture-tube on which the 405-line picture is displayed, and on the right is the camera with the image-iconoscope in which this picture is scanned in the 625-line system.

Two points in the network are particular interest. These are the stations at which the British system (405 lines) is converted into the French (819 line) and the Dutch and German (625 line) systems. The conversion into these two systems was effected by "line-converters" located at Paris and at Breda.

The line-converter at Breda, designed at the Philips laboratories in Eindhoven, was installed in a specially equipped trailer (*fig. 2*); the interior of this trailer is shown in *fig. 3*. The essentials of the line-converter are shown in *fig. 4*. The 405-line picture is displayed on a picture-tube and taken up by a special television camera which scans it in the 625-line system. The principle is thus very simple but precautions are necessary to suppress various disturbing effects. Further details of this apparatus will shortly be published in this Review.

A SHORT LENGTH DIRECT-VIEW PICTURE-TUBE

by J. L. H. JONKER.

621.385.832:621.397.62

The design of cathode-ray tubes for television reception is in a continuous state of development. Progress is usually directed along conventional lines, but occasionally an unorthodox approach yields fruitful results. This article describes a novel approach to the problem of shortening the length of picture tubes; although this development has not been accepted for production purposes, it appears to work surprisingly well.

The original cathode-ray tubes employed in television receivers were no different from those which had for some time previously been used for oscilloscopes. It was not long, however, before these were no longer able to meet the increasingly high standards set for television reception, for which a separate class of cathode-ray tubes came to be developed, known as television picture-tubes (in America: "kinescopes").

As far as these requirements are concerned, the public demand has been (and still is) for the largest possible picture compatible with a receiver of reasonable dimensions and price. As to the dimensions, designers of television sets request that the tube shall be as small (particularly as short) as possible; furthermore, to reduce costs, they prefer to use the lowest possible amount of power for focusing and deflection of the beam. Also, it must be possible to mass-produce the tube in such a way as to ensure uniform characteristics.

In recent years there has been no lack of evidence of the general trend in the development of picture-tubes. Although the tubes remain basically cathode-ray tubes, the dimensions and form have undergone such modification that the requirements referred to above can now be more closely satisfied. Under pressure of public demand, larger and larger screens have been made [1,2]); in the United States, for example, screens have now reached a size of 75 cm diagonal. If it were not for the fact that the ratio of tube length to screen width has also undergone a radical decrease in the meantime, it would have been necessary — even with medium-sized pictures — to make TV cabinets so deep that they would have represented a real obstacle in the average living-room. Thus reduction in the length of picture-tubes has been one of the main objects of the designer in the last few years.

Means of shortening the tube

The length of a tube can be divided roughly into three parts (*fig. 1*), viz. the lengths l_1 of the neck and l_2 of the cone, and the depth l_3 of the screen, which has to be curved to withstand the pressure of the atmosphere.

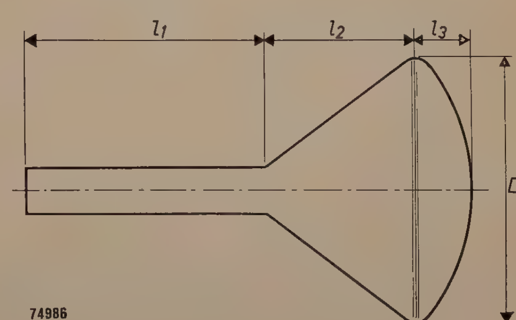

Fig. 1. The length of a television picture-tube is the sum of the neck length (l_1), the cone length (l_2) and the depth of the screen (l_3). D is the diameter or diagonal of round or rectangular screens respectively.

Every effort has been made to reduce the length of each of these parts in relation to the diagonal D; by achieving the maximum compactness of the electron gun and the focusing and deflection coils, it has become possible to effect some reduction in the length of the neck; we shall refer to this again presently. The depth of the screen (l_3) has been reduced in those tubes of which the cone is made of metal instead of glass, as a smaller curvature is then practicable [3]). The most important reduction, however, concerns the cone. Any reduction in the length of the cone will of course be accompanied by an increase in the angle of deflection α, which is

[1]) Round screens are steadily giving way to rectangular screens, thus saving the space occupied by those segments of the round screens which the rectangular picture does not utilize.

[2]) Another method of securing a large image is by projection (see Philips tech. Rev. **10**, 69-78, 1948), but this is not within the scope of the present article.

[3]) J. de Gier, Th. Hagenberg, H. J. Meerkamp van Embden, J. A. M. Smelt and O. L. van Steenis, A steel picture-tube for television reception, Philips tech. Rev. **14**, 281-291, 1953 (No. 10).

the angle between the extreme limits of the deflected beam (*fig. 2*). Step by step this angle has been increased [4]) from 50° to 70° or even 90°, and this has necessitated a greater number of ampere-turns for the deflection; more, in fact, than would be proportional to the increase in the angle α. This results

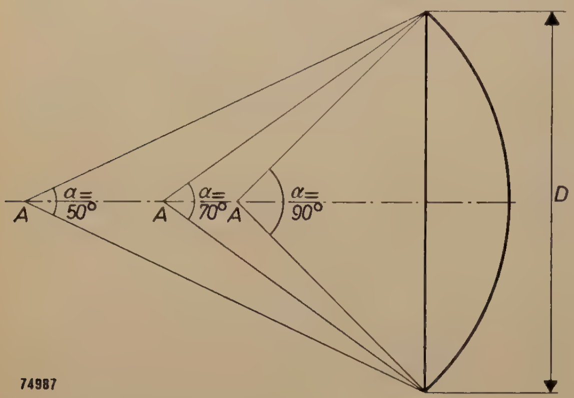

Fig. 2. If the cone be shortened, D being constant, this involves an increase in the deflection angle α. In older tubes this angle was about 50°, but in modern tubes α is 70° to 90°.

from the fact that the point about which the beam pivots when deflected lies in the centre of the deflection coils; if the deflection angle α is increased, the deflection coils must be shorter, as will be seen from *fig. 3*. This means that the electrons are subjected to the deflecting field over a shorter distance, i.e. that an increase in the strength of this field in proportion to α is not sufficient.

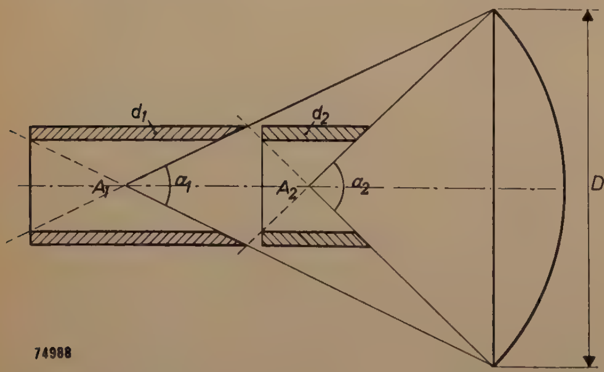

Fig. 3. If the deflection angle be increased from α_1 to α_2, the deflection coils (d_1 and d_2) must be shorter, since the pivoting-point of the beam (A_1, A_2) must always be in the centre of the coils.

In general, a large number of ampere-turns will demand a greater amount of power. However, means have been found to reduce the amount of power required, the more important of these being as follows.

1) So-called economy circuits have been devised for producing the deflection current, whereby a large part of the magnetic energy that accumulates in the deflection field can be recovered and fed back into the supply [5]).
2) Special amplifying tubes (e.g. PL 81 and PL 82) ensuring higher efficiency, have been designed for these economy circuits.
3) Losses in the deflection coils have been reduced, e.g. by the use of Ferroxcube.
4) Economy in the power required for deflection can be achieved in the first instance by reducing the neck diameter of the tube, as far as is compatible with mechanical strength. As will be seen from *fig. 4*, however, a reduction in the thickness of the neck leads to shorter deflection coils. As pointed out, a short coil requires more ampere-turns, and the saving is therefore less than anticipated; but this effect can be largely counteracted by making the cone and neck merge into each other gradually in a certain manner and adapting the deflection coils to the resultant contour [6]).

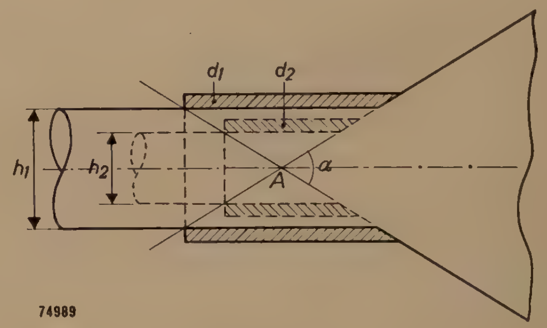

Fig. 4. A reduction in the neck diameter from h_1 to h_2 means that the deflection coils (d_1, d_2), for the same angle α, must be shorter.

However, all these devices do not alter the fact that the amount of power needed for deflection purposes constitutes a limitation on the size of the angle α. And this is not the only limitation. As the angle α is made larger, two further adverse effects become apparent.

As the cone is shortened, i.e. as the pivoting-point of the beam (A, fig. 2) is displaced in the direction of the screen, the surface defined by the focus on deflection will differ more from the surface of the screen, this being accompanied by greater fluctua-

[4]) L. E. Swedlund and H. P. Steiner, Short 16-in. metal-cone kinescope development, Tele-tech **9**, 40-43 and 59-60, Aug. 1950. H. W. Grossbohlin, The design of 90° deflection picture tubes, Tele-tech **10**, 42-44, Aug. 1951.

[5]) See J. Haantjes and F. Kerkhof, Philips tech. Rev. **10**, 307-317, 1949. An article on new economy circuits will later be published in this Review.

[6]) C. V. Bocciarelli, Low-power deflection for wide-angle C-R tubes, Electronics **25**, 109-111, Sept. 1952.

tion in the size of the light spot. In order to limit this defocusing effect as much as possible, the greatest attainable depth of focus of the beam must be aimed at, i.e. the beam should be as narrow as possible [7]).

The other effect experienced with increasing α is related to "pin cushion" distortion. This occurs when the surface defined by the focus does not coincide with the surface of the screen. It is corrected by making the deflection field non-homogeneous — the field being weaker at points away from the axis. However, when the beam passes through an inhomogeneous field, a certain amount of aberration is introduced (which increases with the deflection) as a result of the unequal deflection of different parts of the beam. The narrower the beam, therefore, the less the aberration.

Both these effects can be counteracted only by using a narrower beam. Apart from refinements in the focusing, this can be done only by reducing the beam current (and hence the brightness of the picture), or by increasing the anode potential, which entails higher costs. A practical limit on beam attenuation, and thus also on the angle α, is accordingly soon reached.

The conclusion, therefore, is that any increase in the deflection angle may be accompanied by a useful reduction in the length of the tube, but also involves less desirable effects which in turn have to be remedied. It seems doubtful whether the angle α will be made much larger than 90°, which is already quite large.

Let us now consider the neck of the tube once more. It has been found possible to shorten this to some extent by reducing the length of the deflection coils and, again, by using a shorter electron gun. As regards the latter, two of the components have undergone some modification in the last few years, viz. the focusing system and the ion trap; we shall now take these in turn.

The focusing system

The electron beam can be focused by means of a magnetic lens, an electrostatic lens, or a combination of both. In picture-tubes the first-mentioned method is the more usual.

The reduction in the cone length referred to above has brought the focus nearer to the lens; in other words, the image distance has been reduced and the lens had to be made more powerful. In any shortening of the neck, i.e. reduction in the object distance, the lens must also be stronger. If the magnetic field is generated by a coil, a stronger lens means more energizing power and this is not readily acceptable.

Different ways out of the difficulty have therefore been sought, and one of these consists in combining the magnetic lens with electrostatic pre-focusing: a less powerful lens is then necessary. This has the disadvantage, however, that the stray field of the magnetic lens, at the point where the electrostatic lens in situated, must be negligibly weak; otherwise, if the centering is not perfect, astigmatism occurs. This necessitates a certain neck length, which partially nullifies the reduction in the length already obtained.

A similar difficulty is encountered in wholly electrostatic focusing, to which much attention has been given in recent years with a view to economy in materials [8]). Here it is the stray field of the deflection coils that has to be kept away from the electrostatic lens. A special form of construction is therefore necessary which, in turn, once more increases the length of the neck.

Perhaps the best method is one that makes use of permanent magnets; no energizing is then necessary, and a considerable quantity of copper is saved. Furthermore expensive magnet steels can now be replaced by the non-metallic material Ferroxdure, which contains no scarce materials such as cobalt or nickel [9]).

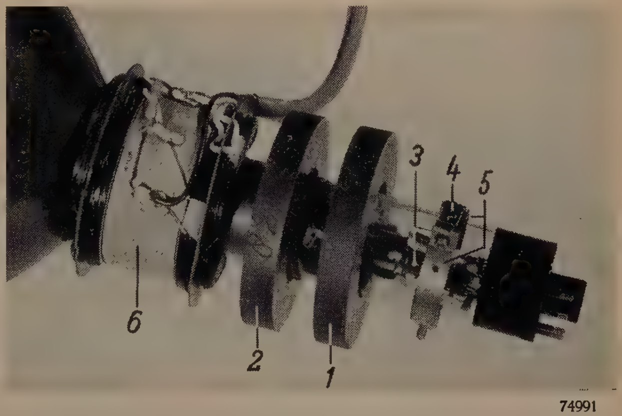

Fig. 5. Picture-tube in which the electron beam is focused by two ring-shaped magnets (*1*, *2*) of Ferroxdure. These rings are magnetized in the axial direction of the tube and are mounted with like poles facing each other. *3* electron gun with ion trap comprising a small steel magnet (*4*) and two pole pieces (*5*). *6* deflection coils.

7) The photograph on p. 368 of this issue illustrates a method of checking the deflection defocussing.

8) L. E. Swedlund and R. Saunders, Material-saving picture tube, Electronics **24**, 118-120, April 1951.
C. S. Szegho, Cathode-ray picture tube with low focusing voltage, Proc. Inst. Rad. Engrs., **40**, 937-939, 1952 (No. 8).
C. T. Allison and F. G. Blackler, A univoltage electrostatic lens for television cathode-ray tubes, Conv. Brit. Contrib. Telev., 1952, art. R8-1333.

9) J. J. Went, G. W. Rathenau, E. W. Gorter and G. W. van Oosterhout, Philips tech. Rev. **13**, 194-208, 1952 (No. 7).

Fig. 5 illustrates the method of focusing by means of Ferroxdure magnets; two flat rings of Ferroxdure are used, these being magnetized in the axial direction and fitted to the neck of the tube with like poles facing each other. The desired configuration of the field is obtained by varying the space between the magnets. The high coercive force of Ferroxdure renders this material very suitable for magnetizing in the direction of the thickness when made in the form of flat rings [10]) and, as we wish to keep this thickness as small as possible, the ring-shaped magnet serves the purpose well.

The ion trap

When the picture-tube is operating, negative ions are produced which originate from residual gases, or from the cathode. Now, since the mass of an ion is very much greater than that of an electron, the field of the deflection coils has very little effect on the ions and, if nothing is done to check them, they all strike the centre of the screen, where chemical action takes place and reduces the luminescence. After a time, a dark spot becomes visible in the centre of the picture.

To prevent this, an "ion trap" is incorporated in the electron gun, and this functions by reason of the circumstance just mentioned, that a magnetic field deflects ions much less than electrons; in an electric field, however, the degree of deflection is the same. *Fig. 6a* illustrates the principle of the ion trap. An electrostatic lens is employed to deflect the paths of the electrons and ions from the axis. Further along, successive deflections by two permanent magnets bring the electron beam back to the axis, whilst the ions strike the second acceleration electrode (the anode), where they are rendered harmless.

A reduction in the length of the neck can be achieved by mounting the first part of the electron gun — viz. the cathode, the control grid and the first acceleration electrode — at an angle with the main axis (fig. 6*b*, fig. 5); this dispenses with the need for one of the deflections, and one of the magnets thus becomes superfluous.

The induction which the ion-trap magnet has to provide is 3.5×10^{-3} to 6×10^{-3} Wb/m² (35 to 60 gauss), and a small steel magnet is generally used (4, fig. 5).

One function of the ion trap, not implied in the name but just as important as the neutralizing of ions, is the means of accurately centering the beam when once the deflection coils and focusing system have been mounted on the tube; in the absence of such centring the oblique passage of the beam through the deflection coils produces distortion of the image. The ion trap magnet thus permits correction of residual errors in the centering.

[10]) See article referred to in footnote [9]), p. 196.

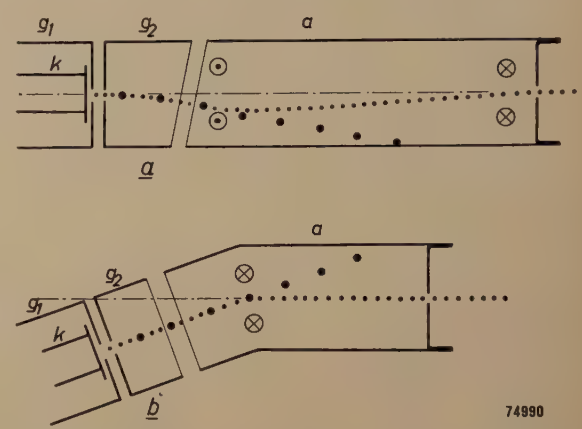

Fig. 6. Old (*a*) and new (*b*) form of ion trap. *k* cathode; g_1 control grid (Wehnelt cylinder); g_2 first acceleration electrode (low positive potential); *a* second acceleration electrode (anode, at high positive potential). Small dots: electrons. Large dots: negative ions.

Action in (*a*): The oblique gap between g_2 and the anode *a* imparts to the field a component perpendicular to the axis of the tube, which deflects the paths of electrons and ions an equal extent from the axis. A magnetic field ⊙ perpendicular to the axis returns the electrons to the axis, but has practically no effect on the direction of the ions, which strike the anode and are thus rendered harmless. A second magnetic field ⊗ opposed to the first, deflects the electrons, so that they move parallel to the axis.

Action in (*b*): k, g_1, g_2 and the first section of the anode *a* are here mounted with their axis at a certain angle to the axis of the tube. Only one magnetic field (⊗) is required to bring the path of the electrons parallel to the axis. The ions are hardly affected by this field and again fall on the anode.

Tube with bent neck

In the foregoing we have outlined some of the obstacles encountered when efforts are made to reduce the length of picture tubes in their present form. Some of these obstacles will doubtless be surmounted in due course, but we will not speculate on such possibilities here. Instead, we shall describe a tube which is considerably shorter than those of the conventional type and which was constructed as an experiment in the Philips laboratories at Eindhoven.

Arising from the effect of the ion trap, in which the beam is given a permanent deflection, the idea was conceived of bending the neck of the tube through an angle of 90° or more and making the beam follow the curve by means of a magnet. In this way the length of the tube can be appreciably reduced (*fig. 7*), the bend serving simultaneously as an ion trap [11]).

[11]) Proc. Inst. Rad. Engrs. **36**, 1485, 1948 (fig. 4) contains an illustration of a tube whose neck is bent slightly, solely to produce the effect of an ion trap.

Fig. 7. Left: Conventional picture-tube (type MW 36-22). Right: Experimental tube with bent neck to reduce the over-all length of the tube.

The most suitable point for the bend is between the focusing system and the deflection coils, the latter being then so constructed that they can pass over the bend; this has, in fact, proved quite practicable.

For curving the beam a small permanent magnet was employed (*figs 8* and *9*) having an induction of roughly 7×10^{-3} Wb/m^2 (= 70 gauss). Adjustment of the position and strength of the field (the strength by means of a magnetic shunt) centres the beam exactly on the axis of the deflection coils in the same way as does the ion trap. To avoid astigmatism, the deflection field must be symmetrical about the plane bisecting the deflection angle, and this is not difficult to achieve. A narrow beam is again an advantage.

In the construction of the experimental tube depicted in fig. 7 we used as many as possible of

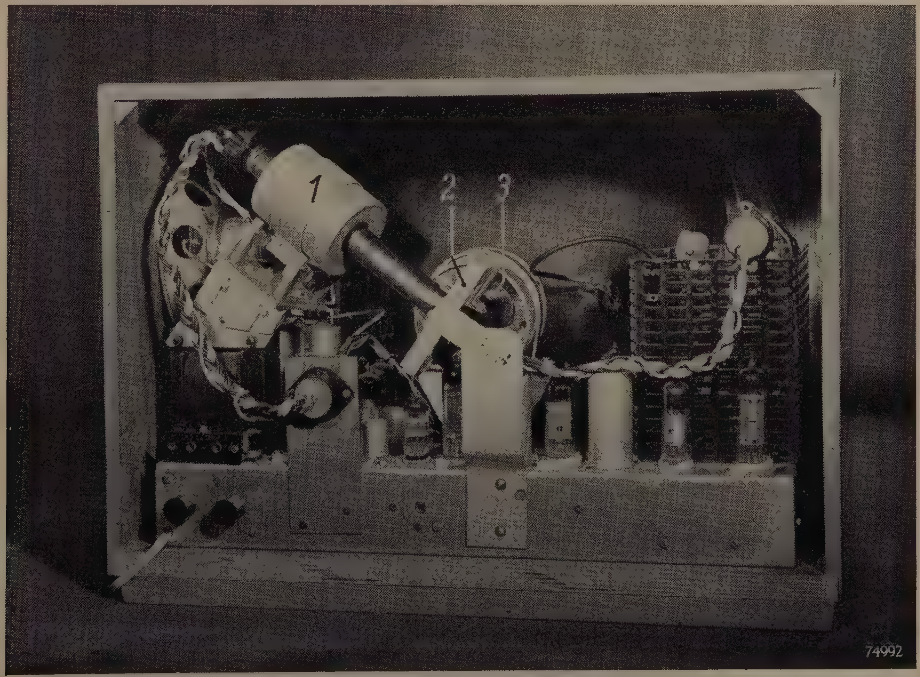

Fig. 8. Tube with bent neck mounted in a television receiver. *1* focusing coil; *2* deflection magnet to make the beam follow the curve in the neck; *3* deflection coils.

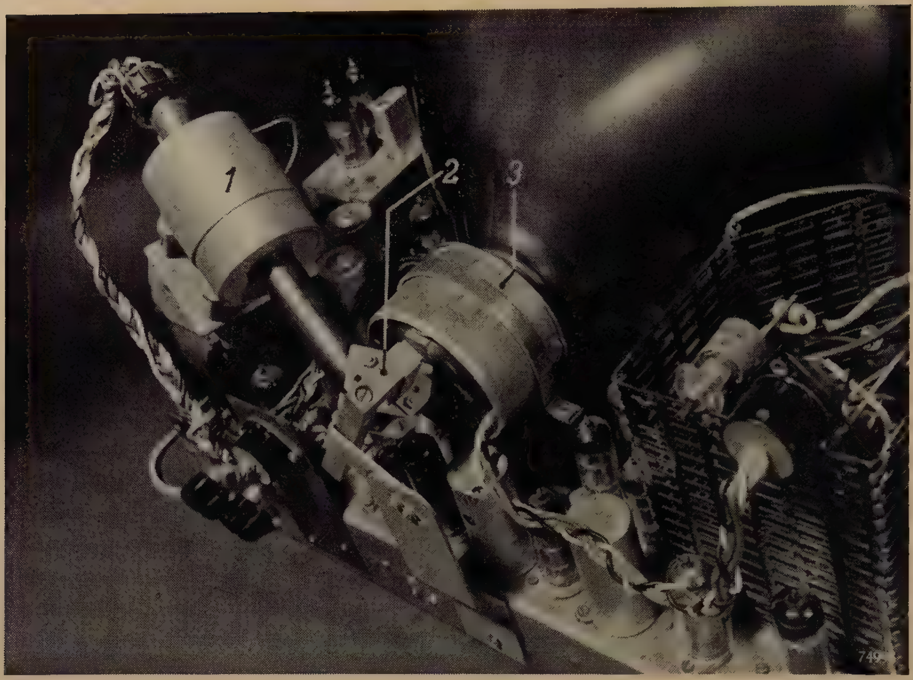

Fig. 9. Close-up of the tube neck and deflection coils.

the standard components of tube MW 36-22, viz. the screen (25 cm × 32.5 cm), the cone, the electron gun (without ion trap) and the deflection coils. The deflection angle (α) is the one usually employed, viz. 65°; the normal deflection power is accordingly sufficient. Only the focusing coils and the permanent magnet were specially made.

As the length of the bent part of the neck in no way affects the depth of the television cabinet, it can, if desired, be longer than that of a tube with straight neck; the neck of the experimental tube is, in fact, longer as this ensures better focusing: the distance from object to lens is greater, the magnification correspondingly less and the light spot smaller. At the same time, a long focusing coil can be used, the diameter of which will then be smaller. By bending the neck through an angle larger than 90° we ensure that the focusing coil lies within the over-all length of the tube, so that only this length determines the depth of the receiver cabinet, not the tube length plus the thickness of the coil.

The direction in which the neck is bent is immaterial; in this tube the direction lies in the plane of the diagonal of the screen, for reasons of available space in the cabinet (fig. 8).

A neck bent in the horizontal plane also has certain advantages. Such an arrangement would facilitate supporting the focusing coil and the permanent magnet on the chassis. A second advantage is related to the fact that, when a permanent magnet is used for bending the beam, the picture is displaced slightly on the screen when the high tension varies. With the neck of the tube bent horizontally this displacement would also be horizontal, and this would not be so noticeable as diagonal displacement, as is now the case. It may be added that such displacement can be entirely avoided by employing a stabilized high tension supply, the output of which is constant in spite of variations in the mains voltage or in the load [12]).

Using this tube a television receiver was constructed (*fig. 10*), the over-all dimensions of which were as follows:

width	20″,
height	$14\frac{1}{2}$″,
depth	$13\frac{3}{4}$″,

i.e. no larger than a medium-sized radio receiver.

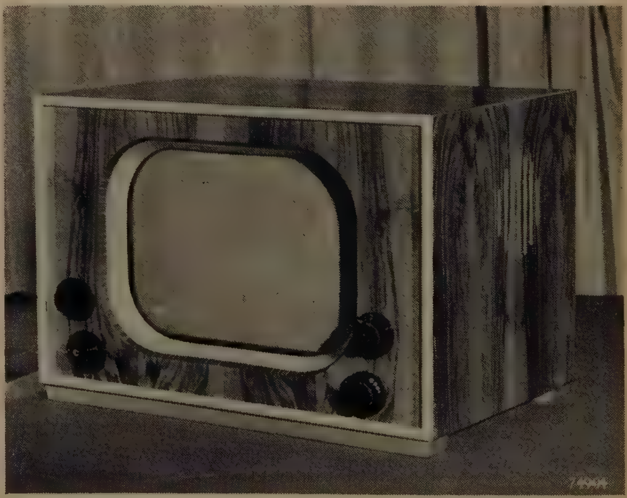

Fig. 10. Front view of the receiver illustrated in figs 8 and 9. The loudspeaker is at the side.

[12]) See for example, J. J. P. Valeton, Philips tech. Rev. 14, 21-32, 1952 (No. 1).

The height has been kept down by mounting the loudspeaker at the side of the cabinet instead of below the tube, a method that is quite often employed (see also fig. 8).

It cannot be pretended that the bent-neck tube is more than an experimental idea, and it should not be expected that such tubes will be put into production for the present. Though practical application may be delayed, however, it was considered that publication of a description of this laboratory model was not inopportune.

Acknowledgements for their co-operation in the design of both tube and receiver are due to Messrs. Dammers, Diemer, Neeteson and De Weyer.

Summary. Arising from the demand for larger television pictures, the development of direct-vision tubes has been directed towards larger and larger screens. This has of course resulted in greater tube lengths and, as this length determines the depth of the receiver cabinet, television sets have shown a tendency to become rather unwieldy. In recent years, therefore, every effort has been made to reduce the length of the tube in relation to the picture dimensions. Some of the methods employed are outlined, viz. shortening the cone (accompanied by wider deflection angle), reducing screen curvature (as in the metal-coned tube), focusing by Ferroxdure permanent magnets, and the use of simplified ion traps.

A description follows of an experimental tube whose neck is bent through 90° or more, resulting in an appreciable reduction in the over-all tube length. A permanent magnet is used to make the beam follow the curve in the neck. A television set was built for this tube, the depth of which was only 13¾″ (less than the diagonal of the screen); the width and height are 20″ and 14½″ respectively. The deflection angle is the conventional 65 degrees.

CHECKING THE LUMINOUS SPOT IN CATHODE-RAY TUBES

Photograph Walter Nürnberg

The quality of a television picture is largely dependent on the sharpness of the luminous spot which builds up the image, line by line, on the screen of the cathode-ray tube — it depends, thus, on the accuracy with which the electron beam is focused to a point on the screen. The above photograph shows the inspection of spot size and shape during the manufacture of picture tubes. The tube is set up under normal operating conditions and fed via the grid with a series of voltage pulses in such a way that a raster of spots is produced on the screen. The voltage pulse magnitude is so chosen that the amplitude of the pulsed electron beam current obtains a particular value e.g. 100 μA. After careful adjustment of the focusing current, a microscope is used to determine whether or not the spots are small enough and sufficiently circular.

A BATTERY-OPERATED GEIGER-MÜLLER COUNTER

by G. HEPP *).

621.317.7:539.16.08:621.353:621.387.424

The development of counter tubes during the last few years has been characterized by their adaptation to specific problems that have arisen in scientific and technical applications. Today, in consequence, counter tubes are employed for a variety of purposes in quite simple equipment. In many of these applications small and conveniently portable units are desirable, and the following is a description of a Geiger counter for battery operation which, by reason of its small dimensions and low weight, will prove a valuable instrument in a wide range of uses.

The widening field of applications of radio-active substances and equipment that emits ionizing radiations (X-ray tubes, cyclotrons, betatrons etc.) is creating a growing demand for simple, easily-handled instruments for detecting and measuring radio-active radiations. Geiger counters commonly form the basis of instruments for this purpose. Such instruments are used amongst other things for locating radio-active ores and for the protective monitoring of radiations in laboratories. An example of a more technical application is the measurement of liquid level by means of a float containing a radio-active substance, the height of the float being ascertained with the aid of a counter. (The advantage of this method over numerous others is that it is not necessary to make any connections through the wall of the vessel.)

In the following, a description is given of a Geiger counter suitable for battery operation. As the current consumption is very low, small batteries can be used, of the kind employed in hearing aids, and the instrument is accordingly quite compact. It is fitted with a Geiger-Müller counter tube in which the quenching gas is a halogen [1]).

Working principle

The measurement of radiations by means of counter tubes generally takes place along the following lines. Every discharge in the counter tube produces a voltage pulse across a resistor in series with the tube, and these pulses are fed to a circuit which converts them into current pulses whose amplitude and duration are independent of the intensity of the discharge in the tube. (The discharge intensity is dependent on the voltage applied to the tube, as also on the characteristics of the tube itself. The first of these drawbacks could be overcome by stabilizing the supply voltage, but this would mean too great a drain on the batteries. Furthermore, in view of the second disadvantage, pulse-shaping is in any case generally preferable.)

The shaped current pulses are applied to a capacitor which is charged across a resistor, the discharging current being then measured with a D.C. measuring instrument. Seeing that the discharges in the counter tube occur at random, the capacitor charging current will be subject to fluctuations. However, if the time constant (RC) of the capacitor C and the discharging resistance R, is large compared to the average time between pulses, the charging current will be sufficiently smoothed to ensure a meter deflection independent of the fluctuating pulse frequency.

The supply voltage for the counter tube can be obtained in different ways, namely:

1) from a battery capable of producing the required high tension. This involves heavy and bulky equipment, with fairly high running costs for battery renewals.
2) From a vibrator operated on a low-tension accumulator. Such vibrators require a considerable amount of power to maintain the motion of the contact springs.
3) From an oscillator, fed from the same battery as that used for the pulse-shaping valve. At first sight, even this method would not appear to be very attractive, since it necessitates the use of an additional valve, with consequent increased load on anode and heater batteries. However, it is possible to eliminate both these objections by employing only one valve for the two functions concerned; the oscillator then operates only at those moments when a discharge occurs in the counter tube. This arrangement ensures very low anode current consumption and consequently a longer battery life.

*) Temporarily with Centro Brasileiro de Pesquisas Fisicas, Rio de Janeiro, Brazil.

1) See N. Warmoltz, Geiger-Müller counters, Philips tech. Rev. **13**, 282-292, 1952.

Economy circuits for a Geiger counter

The principle on which the Geiger counter works is described with reference to *fig. 1* which shows the simplified circuit diagram. The counter tube *GM*, in series with a resistor R_1 is connected in parallel

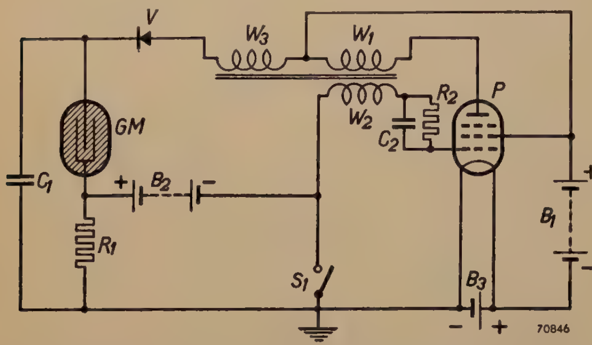

Fig. 1. Circuit diagram of battery-operated Geiger counter. *GM* Counter tube, *P* pentode, *V* rectifier element, B_1 H.T. battery, B_2 blocking battery, B_3 filament battery, S_1 starting switch.

with a capacitor C_1 which, under ordinary conditions, carries a sufficiently large charge to initiate an independent discharge of short duration in the counter tube on the arrival of an ionising particle. With each such discharge in the tube the capacitor C_1 loses some of its charge, and the voltage accordingly drops each time. After each discharge, the capacitor is re-charged to the requisite voltage via a rectifier element *V*; the current pulse which is passed to the capacitor through the rectifier is supplied by a pentode *P*, connected as a blocking oscillator [2]).

When switch S_1 is closed (so that the battery B_2 no longer affects the oscillator), current pulses occur at regular intervals in the anode circuit of the valve *P*. The time elapsing between pulses is governed mainly by the values of the capacitance C_2 and the resistor R_2. The transformer, of which W_1 is the anode winding and W_2 the grid winding, includes a third winding W_3, so connected that each current pulse in the anode circuit induces a pulse in W_3 which charges C_1 through *V*.

The function of the battery B_2 is to prevent the occurrence of current pulses in *P* when no discharges are taking place in the counter tube. This battery lowers the potential of the grid to a level with respect to the cathode such that oscillation cannot occur (with switch S_1 open). A discharge in the counter tube produces a voltage pulse across the resistor R_1 which momentarily neutralises the voltage from the battery B_2, and permits P to pass a current pulse.

As the valve *P* does not work continuously as a blocking oscillator, but only at those moments when the counter tube operates, the battery B_1 is not under continuous load and the circuit accordingly consumes very little current (the filament battery B_3 is of course loaded continuously, but the filament current of the valves used is only 13 mA at 1.5 V).

Since a discharge in the counter tube is essential for the occurrence of a charging pulse from the pentode and, as the discharge is possible only when the capacitor C_1 carries a certain charge, the circuit cannot start to function spontaneously. For this reason switch S_1 is provided to isolate the battery B_2 from the grid circuit and temporarily remove the block on the pentode *P*. When the instrument is first switched on, therefore, this switch must be closed for a moment, to enable the pulses to charge the capacitor C_1 to the required potential. When switch S_1 is opened again, each discharge taking place in the counter tube causes the charge on C_1 to be replenished in the manner already described, since the charge per pulse that can be applied to C_1 through *V* is greater than that utilized by the counter tube per count.

Unfortunately, however, the counter tube is not the only source of loss in charge of C_1. For the lowest count rate, i.e. the "background" count (number of discharges per unit time in the absence of a specific ionizing event), the losses due to leakage in the capacitor itself, in the rectifier *V* and in the counter tube are much more significant. Consequently, unless C_1 receives an adequate charging current, the voltage across it will drop below the working voltage of the counter. As the instrument must continue to operate when once switched on, the leak currents must therefore be kept as low as possible. For this reason *V* cannot be a selenium or germanium rectifier, as the leakage of these is too high. A diode in the circuit shown in fig. 1 would necessitate an additional filament battery, but a circuit which dispenses with the need for this is depicted in *fig. 2*. Here, instead of a positive voltage at the anode of the counter tube as in fig. 1, a negative voltage is produced at the cathode; in this circuit use is made of a diode *D* included in the envelope of the pentode. The capacitor which is charged by the current pulses is once more denoted by C_1. The voltage pulses produced across the resistor R_1 are applied to the control grid of the pentode *P* through another capacitance C_2, so that

[2]) For the operation of a blocking oscillator see, for example, D. Goedhart and G. Hepp, Carrier supply in an installation for carrier telephony, Philips tech. Rev. **8**, 137-146, 1946, and P. A. Neeteson, Flywheel synchronisation of saw-tooth generators in television receivers, Philips tech. Rev. **13**, 312-322, 1952 (No. 11).

the pentode is cut off each time the counter tube has functioned. A special feature of the circuit illustrated in fig. 2 is that the potential occurring across the counter tube is the sum of the potential developed across C_1 and that of the battery B_1; it is thus not necessary for C_1 to develop the whole of the voltage required for the counter tube. If the voltage needed were, say, 360 V and that of the battery B_1 45 V, only 315 V would have to be developed across C_1. A difficulty in connection with this circuit is that the leakage of capacitor C_1 is in part due to the finite insulation resistance of winding W_3 of the transformer; it has been found impossible in practice to attain a sufficiently high insulation to entirely eliminate this leakage.

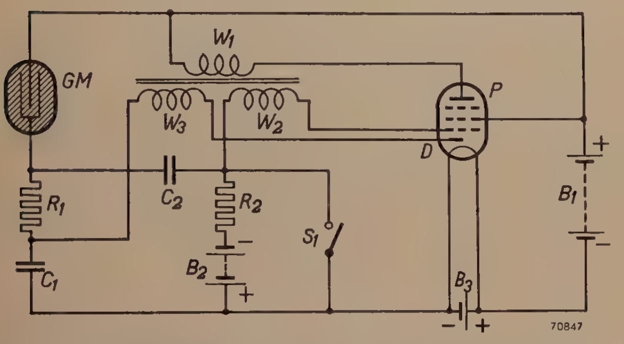

Fig. 2. Circuit diagram in which the rectifier shown in fig. 1 is replaced by a diode D included in the envelope of the pentode.

A simple calculation will determine the permissible order of size for the leak resistance. At a voltage of 360 V and with 1000 pulses per second, the mean value of the current flowing in the counter tube is about 1 μA. A charge of 10^{-9} coulombs per pulse is therefore taken from the capacitor. If we denote the number of pulses per second by N, the counter tube will in that time consume a charge equal to:

$$N \times 10^{-9} \text{ coulomb}.$$

Further, if we denote the leak resistance of capacitor C_1 by R_1, the leakage per second with 315 V across the capacitor will be:

$$\frac{315}{R_l} \text{ coulomb}.$$

With a very large number of counts per second taking place in the counter tube, the amount of the charge that the tube will take from the capacitor will be high compared with the loss due to the leakage current. For a low counting rate, however, the leak current gains in significance, since the charging pulses are proportional to the number of discharges taking place in the counter tube. Now, it is essential that, even at the lowest count rate, namely the background count, there will be sufficient surplus charge in the capacitor C_1 to compensate the leakage current.

Let I denote the maximum average anode current relating to a single charging pulse, T the duration of the pulse, and W_1/W_3 the ratio of primary to secondary (W_3) turns. The maximum charge that can be supplied to C_1 per second is then:

$$NI \frac{W_1}{W_3} T \text{ coulomb}.$$

To ensure that the potential across C_1 will not drop, this charge should be equal to or greater than

$$(N \times 10^{-9}) + \frac{315}{R_l} \text{ coulomb}.$$

If $I = 2$ mA, $T = 10^{-4}$ sec and $W_1/W_3 = 1/10$, $N \times 10^{-9} \ll NI\,(W_1/W_3)\,T$, so that:

$$NI \frac{W_1}{W_3} T \geq \frac{315}{R_l}.$$

Hence, if one discharge occurs every 5 sec ($N = 0.2$) in the "background", the leak resistance is required to be

$$R_l \geq 10^{11}\,\Omega.$$

A circuit which imposes less strict requirements on the insulation resistance of the transformer is depicted in *fig. 3*, in which the capacitor C_1 is connected between winding W_3 of the transformer and the diode anode. The other end of W_3 is earthed. In this kind of circuit great care must be taken to keep the capacitance between C_1 and the chassis as small as possible. As will be seen from fig. 3, this stray capacitance is in parallel with the self-capacitance of winding W_3, these together being designated as C_p in the diagram. In order to raise the anode voltage of D to a level high enough to permit the charging of C_1, the height of the pulse across W_3 must be at least equal to the potential across C_1. C_p, which is charged repeatedly to this potential, must therefore be as small as possible. After each pulse, C_p discharges via the transformer winding. This charging energy is lost, at least in so far as the charging of C_1 is concerned. (It will be shown later, however, that this energy can be used for another purpose, i.e. to produce a deflection in the indicating instrument.)

The permissible magnitude of C_p can be estimated as follows: with the above-mentioned values of T, I and W_1/W_3, the maximum available charge per pulse is 2×10^{-8} coulomb.

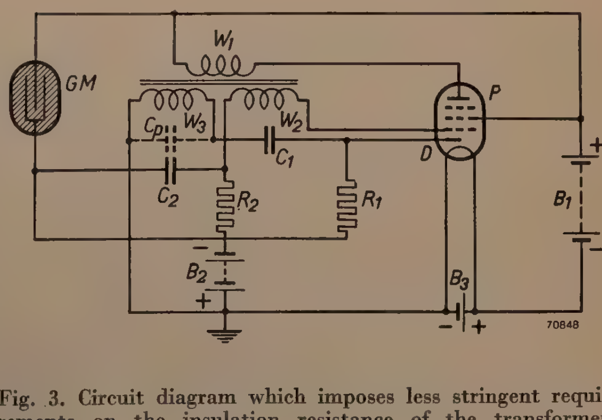

Fig. 3. Circuit diagram which imposes less stringent requirements on the insulation resistance of the transformer winding W_3.

If C_1 and C_p are charged to a potential of 315 V, and not more than 10% of the charge is to be applied to C_p, the value of C_p must be such that:

$$315\ C_p \leq \frac{2 \times 10^{-8}}{10}$$

or:

$$C_p \leq 6.3\ \text{pF}\ .$$

It has been found in practice that this can be accomplished by employing a special winding and insulating technique for the transformer, and by careful positioning of the components.

In order to keep the capacitance between C_1 and the chassis as low as possible, the physical dimensions of this capacitor must be limited, i.e. the value of the capacitor will also be low. A low capacitance is also desirable for another reason, viz that a short starting time is required, this being the time that switch S_1 must be kept closed for the capacitance C_1 to charge up (see figs 1 and 2). This period should preferably be limited to a fraction of a second and, of course, the lower the capacitance of C_1, the shorter this will be. On the other hand, if the capacitor is too small, there is every risk that, with only a few pulses arriving within a given time, the voltage across C_1 will drop so much between pulses that the counter tube is unable to function, thus interrupting the working of the circuit. A sufficiently high value of C_1 is therefore used to minimise the chance that the circuit will cease to function, even with the smallest number of pulses (background) likely to occur in practice. The probability of failure due to this cause may be estimated numerically in the following way.

Suppose E_c to be the minimum voltage across the counter tube which makes the counting pulses large enough to remove the block on the oscillator. Taking E_n as the normal tube voltage, we introduce the quantity

$$p = \frac{E_c}{E_n}\ .$$

If N_o is the number of pulses per second in the background and k the product of the capacity C_1 and the leak-resistance R_1, then the time T_c necessary for the voltage E_n to decrease to the value E_c is derived from the relation:

$$E_c = E_n\, e^{-\frac{T_c}{R_1 C_1}} = E_n\, e^{-\frac{T_c}{k}}\ ,$$

from which

$$\frac{T_c}{k} = -\ln \frac{E_c}{E_n} = -\ln p.$$

The chance that a given pulse is not followed within this time by a following pulse is

$$e^{-N_o T_c} = e^{kN_o \ln p} = p^{kN_o}.$$

Of the $N_o T$ pulses occurring in a time T there are thus $N_o T p^{kN_o}$ which are not followed within the time T_c by another pulse. If the value of this expression is small in relation to unity, it gives at the same time the probability that the critical voltage E_c will be reached within the time T.

It is tacitly assumed in this connection that the occurrence of a single pulse is sufficient under all circumstances to bring the voltage back to the value E_n. If, however, more than one pulse is necessary, the same conclusion is valid provided that the number of pulses required is relatively small.

If, for example, $k = 1000$ sec, $p = 0.85$, and $N_o = 1/5$, then during a time T of one hour the chance that the voltage drops below the value E_c is:

$$\tfrac{1}{5} \times 3600 \times (0{\cdot}85)^{200} = 4{\cdot}5 \times 10^{-12}.$$

Stumpers [3]) has discussed this problem under the assumption that no limitation of the capacitor voltage is occasioned by the pulse amplitude, and that each pulse produces the same charge in the capacitor.

Finally, it should be mentioned that exceptionally high requirements are demanded for the self-inductance coefficient of the transformer windings. This must be very high in order to limit the magnetisation current. An estimate of the value required can be made with the aid of an equivalent transformer circuit.

This consists of an "ideal" transformer with windings of infinite self-inductance (leakage inductance and losses being thereby disregarded), and a coil, in parallel with one of the windings, whose self-inductance is equal to that of the winding of the actual transformer. *Fig. 4* shows an equivalent circuit of this kind, based on a part of fig. 3. In this figure T represents the ideal transformer; the current I_3 is equal to $I_1 W_1/W_3$. The pentode P is represented by a switch S_p, by means of which the current I_1 is periodically applied. Suppose that $I_1 = 2$ mA and that $W_1/W_3 = 1/10$; then $I_3 = 0.2$ mA. Immediately the circuit is closed, this whole current will flow in the capacitor C_1 as well as through the diode, but an increasingly large part of I_3 will also flow in the coil of which the self-inductance L_3 is the same as that of W_3 (fig. 3). If the voltage across C_1 is 315 V, the time derivative of the current in L_3 will be determined by:

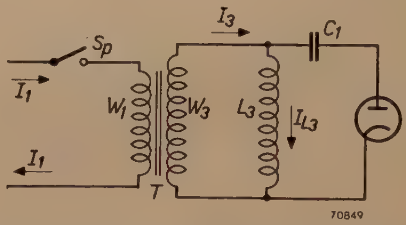

Fig. 4. Equivalent circuit diagram of the transformer. The pentode is represented by a source of current I_1 and a switch Sp.

$$L_3 \frac{dI_{L3}}{dt} = 315.$$

Suppose now that during the period of one charging pulse (10^{-4} sec) the current I_{L3} is not to exceed 20% of the total available current I_3 (0.2 mA), so that not more than about 10% of the current is wasted; dI_{L3}/dt must then be not less than:

$$\frac{0.04 \times 10^{-3}}{10^{-4}} = 0.4\ \text{A/sec},$$

hence L_3 should be greater than $315/0.4 \approx 800$ H.

[3]) J. F. H. M. Stumpers, Philips Res. Rep. 5, 270-281, 1950.

The meter circuit

As explained on p. 369 the mean anode current flowing in the pentode P is a measure of the average number of counts performed by the counter tube per unit of time. The most obvious method of measuring the pulse frequency is therefore to connect a meter with a series resistor, in parallel with a capacitor in the anode circuit of P. Both the meter and the series resistor then occasion a certain voltage drop, however, necessitating a battery B_1 of higher voltage and involving a higher battery drain by the whole unit.

It is found that a lower consumption of current can be achieved by means of another circuit, as shown in *fig. 5*. Here the milliammeter M with resistor R_m

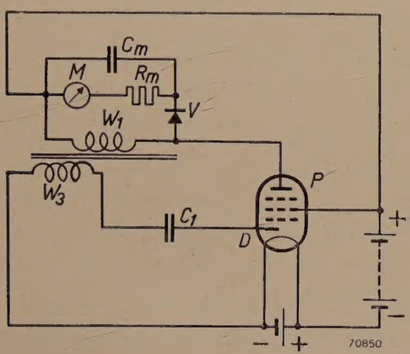

Fig. 5. A method of connecting the meter which is an improvement on a meter placed directly in the anode circuit of the pentode.

is in parallel with a capacitor C_m, this circuit, together with a rectifier V, being connected in parallel with the primary winding W_1 of the transformer. Each time that the pentode P passes a current pulse, a damped oscillation occurs in the oscillatory circuit formed by winding W_1 and its own self-capacitance. Now, the polarity of the rectifier element V and diode D is such that, in the first half of the first cycle, the capacitor C_1 is charged, and, in the second half of the same cycle, the capacitor C_m. During the complete cycle effective use is thus made of almost all the available energy.

Apart from this more effective utilization of the energy, this circuit also ensures greater stability of the whole instrument than if the meter were included directly in the anode circuit of P, for, if the oscillation in W_1 were allowed to decay freely, there would be some risk that the second period would assume the function of a count. The pentode would thus be triggered, whereupon the process would be repeated, and the number of anode current pulses would become independent of the number of counts. By contrast, if practically all the energy is utilized during the first period in the manner just described, the amplitude of the decaying oscillation is very small and all risk of the above-mentioned undesirable consequences is eliminated.

The circuit consisting of the diode-pentode may be regarded as a D.C. transformer fulfilling the same function as the vibrator unit in car-radio sets. The voltage supplied by the battery B_1 is "transformed" up to the necessary value for the counter tube; this voltage is mainly dependent on the battery voltage and the "transformation ratio" and adjusts itself automatically to roughly the required value, irrespective of the number of pulses per second. *Fig. 6* shows the results of measurements that confirm this. The current flowing in the counter tube and the diode are each plotted as a function of the voltage on the counter tube (in this case externally applied), for different numbers of counts. The voltage on the tube of course adjusts itself to the value at which these currents are equal. It is seen from fig. 6 that for a variation in the number of counts from 50 to 600 the voltage on the counter tube varies only between 385 and 360 V.

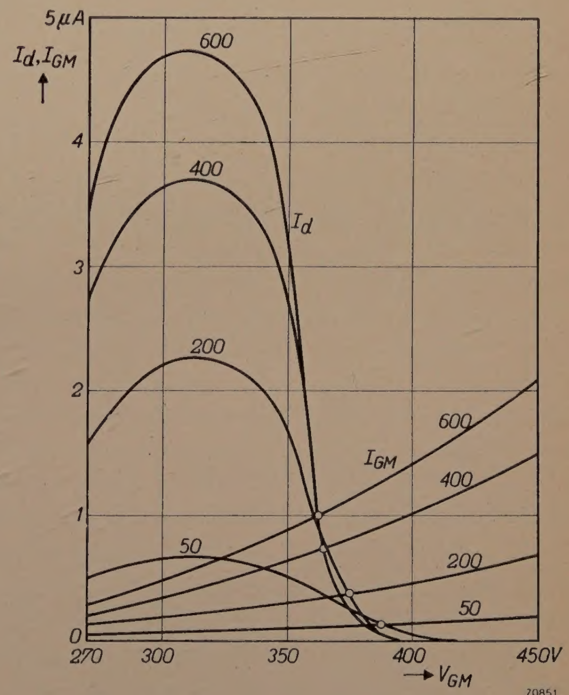

Fig. 6. The diode current I_d and counter tube current I_{GM} as a function of the (externally applied) voltage V_{GM} on the counter tube. The values shown on the curves indicate the number of counts per second. The voltage across the counter tube adjusts itself to the point of intersection of the relevant curves; this voltage is seen to be only slightly dependent on the number of pulses per second.

Final design of the apparatus

The circuit in the ultimate design differs from the diagrams given above in some respects; this final

circuit is depicted in *fig. 7*. The principal difference compared with figs. 1, 2 and 3 is that the transformer does not include a separate winding for the feed-back and rectification; winding W_2 is employed

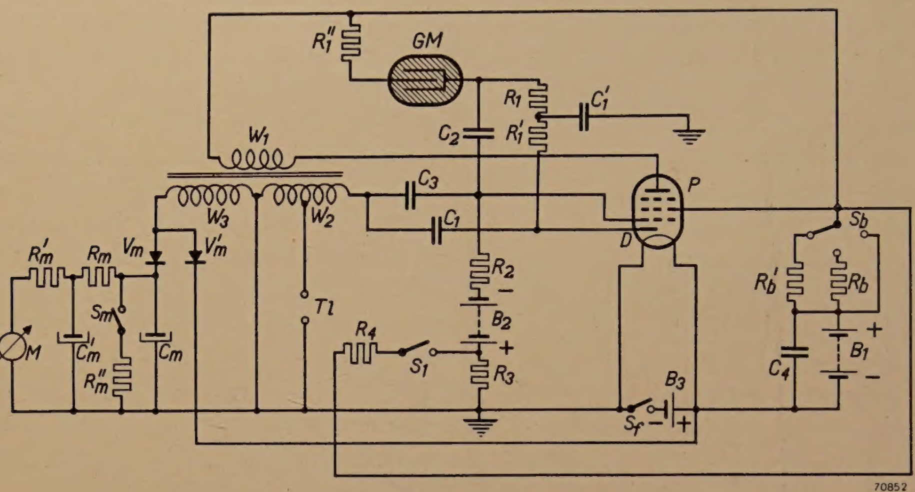

Fig. 7. Circuit of the Geiger counter for battery operation, with certain less important details omitted.

for both purposes. Here, W_3 serves only for connection of the measuring instrument (through the rectifier V_m). The necessary delay for the meter is provided by a filter which is in two sections. A switch S_m, by means of which a resistor R_m'' can be placed in parallel with the capacitor C_m, enables the sensitivity to be adjusted to one of two values, viz. 40 and 600 counts per sec, for full deflection. The filter is made in two sections, because otherwise, when the switch S_m is closed, the low value of R_m'' would result in too small a time constant.

In addition to V_m, a second rectifier element V_m' is included. In most cases this element is inoperative because the low tension battery B_3 supplies it with a certain bias. V_m' functions only when the unit is employed for very intense radiation while still set for maximum sensitivity. In such cases there may be some risk that, during the decay of the oscillation in the circuit of W_3, the second cycle would assume the function of the counter tube, as mentioned above. Each pulse in the anode circuit of P would then be automatically followed by another, and this condition would be maintained even if the intense radiations were no longer present. Upon removal of the unit from the area of such radiations, then, the meter would continue to give full deflection, and this might well lead to confusion between one measurement and another. The rectifier cell V_m' is therefore included in order to limit the voltage on C_m and at the same time to prevent, during the decay of the oscillation, the voltage across W_3 from reaching such a value that the conditions described can arise. In addition, protects V_m' the meter M from overloads in the presence of intense radiations.

Fig. 8 shows the voltage waveform in the transformer windings resulting from a pulse, in the absence of, and with the rectifier element V_m' (figs 8*a* and 8*b* resp.). The flattening of the first half-cycle is due to the charging of the capacitor C_1. If now the voltage rises again during the second cycle to nearly the same value as in the first cycle (fig. 8*a*), the danger mentioned above arises that the block will again be removed from the oscillator. The element V_m' limits the second half of the first cycle, resulting in the waveform of fig. 8*b*. The damped oscillation now has a much smaller amplitude.

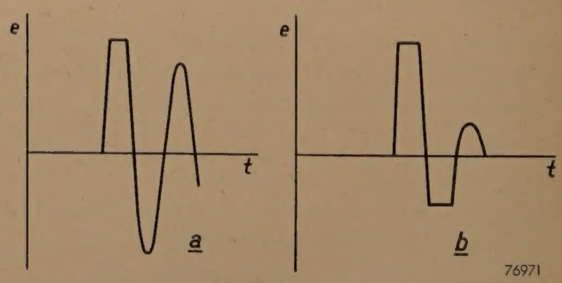

Fig 8. Voltage (*e*) waveform in the transformer windings resulting from a charging pulse, as a function of time (*t*); *a*, in the absence of the rectifier element V'_m, and *b* with V'_m connected.

The rectifier element V_m' could also be connected to the upper side of the capacitor C_m, but in this case, due to the leak in V_m', battery B_3 would give rise to a small reverse-current through this element, resulting in a permanent deflection in the indicating instrument. By connecting V_m' as shown in fig. 7 this current flows through the winding W_3 to the chassis.

The feed-back necessary for the pentode to supply anode current pulses, is produced by a capacitor C_3. It could also be obtained without this capacitor, viz. across C_1, R_1 and C_2 (if R_1' and C_1' were not provided), but it was found that a capacitor connected in the manner shown ensures shorter duration of the anode pulses, this being advantageous when the number of counts per unit of time is high. In order to avoid feed-back across C_1, R_1 and C_2 as well, a resistor R_1' and capacitor C_1' are included.

It is very important that the duration and wave-form of the anode current pulses shall be as constant as possible in all circumstances, for these determine the calibration of the meter. The latter is affected considerably by the voltage supplied by the anode battery B_1; since this voltage drops during the life of the battery, two resistors R_b and R_b' are included for switching in series with the battery for "high" and "average" battery output. The values of these resistors are such that on "normal" battery output a voltage-drop occurs which is just sufficient to reduce the anode and screen grid supplies to the required values during the pulse. The strength and duration of the pulses are thus rendered independent of the value of the battery voltage. A capacitor C_4 is connected in parallel with B_1 to prevent the increasing internal resistance that accompanies ageing of the battery B_1 from affecting the pulse duration.

The instrument is switched on by momentarily closing the switch S_1; a positive voltage, obtained from the potential divider R_3-R_4 across the battery B_1, then compensates the voltage from the battery B_2.

When the number of ionising particles entering the counter tube per unit of time is too small to be indicated by the instrument, the current pulses can be rendered audible by means of headphones; the terminals provided for this purpose are denoted by Tl in the circuit diagram [4]).

Fig. 9 depicts the Geiger counter in use and *fig. 10* shows the instrument with cover removed. The weight is 520 g (approx. 18 ozs.) and the dimensions are $17 \times 4 \times 10$ cm (approx. $7'' \times 2'' \times 4''$). Small batteries of the kind supplied with hearing-aids are used. A knob with 6 positions is provided on the front panel for operating the switch S_1, the sensitivity switch S_m and the filament heater switch S_f. The position of the knob at which S_1 is closed is immediately next to the "off" position. In general, rotation of the switch through the "start" position to one of the other settings is sufficient to put the instrument into operation. The meter can also be employed to check the voltages of the filament battery B_3 and the H.T. battery B_1, these operations being also effected by rotating the knob to the appropriate positions;

67065

Fig. 9. The Geiger counter in use. The illustration reveals the fact that the very weak radio-active radiations from the luminous dial of a wrist watch can be registered.

the associated circuits are omitted in fig. 7 for the sake of clarity. Resistors R_b and R_b' are placed in circuit with the aid of a separate switch, and two marks on the meter indicate the anode voltages at which this switch should be set, in accordance with similar symbols.

Calibration of the instrument

As already mentioned, the instrument has two scales, one for 0-40 and one for 0-600 counts per sec. Calibration of these scales would appear to be simpler than it actually is; it might be supposed that it is only necessary to connect a counting device to the instrument in order to be able to count the number of pulses per second for each deflection of the meter. For a sufficiently small number of counts this would certainly be possible, but, with an increase in the number of ionising particles, the number of discharges in the counter

[4]) In exploration work it is often more convenient to use headphones, as the eyes do not then have to be kept on the meter.

tube drops below the number of particles entering, seeing that the "dead time" of the counter tube as well as that of the blocking oscillator then become more and more significant (see article referred to in footnote [1])). If this dead time were known, it would be a fairly simple matter to compute the number of ionizing particles from the number of discharges per second, but the dead time is not usually known. More accurate calibration is ensured, therefore, by measuring the number of ionising particles with an instrument whose dead time is negligibly short.

By the method actually adopted use is not made of individual particles or photons, but of pulsed X-rays, the pulses being of such short duration that each produces only one count, although the number of photons per pulse is very high.

The Geiger counter is placed in the vicinity of an X-ray tube fitted with a grid which is biased so that normally no anode current flows: positive pulses applied to this grid result in anode current, thus producing pulsed radiation. The frequency of the voltage pulses for the grid varies arbitrarily, the pulses being derived from a radio-active substance emitting α-particles, placed in the neighbourhood of a scintillation counter. In this apparatus voltage pulses are produced each time an α-particle strikes a zinc sulphide layer mounted in the unit; these pulses are amplified and then applied to the grid of the X-ray tube, as well as to a "scaler" which counts the number of pulses, or to a counting-rate meter. The dead time of these instruments is so short that it is of no significance in the measurements. The number of pulses per second can be controlled by varying the distance between the radio-active substance and the scintillation counter, thus providing a ready means of calibrating the Geiger counter.

The fact that only the number of "pulses" of X-rays (or of artificially produced radiations in general) emitted for very short periods of time is counted, and not their intensity, is the main reason why calibration in roentgens per hour is not provided. In certain cases it would otherwise be possible to conclude quite wrongly from the deflection that a given radiation were harmless. Another reason why it is not practicable to calibrate in roentgens per hour is that the sensitivity of counter tubes to γ-rays is dependent on the particular wavelength of these rays.

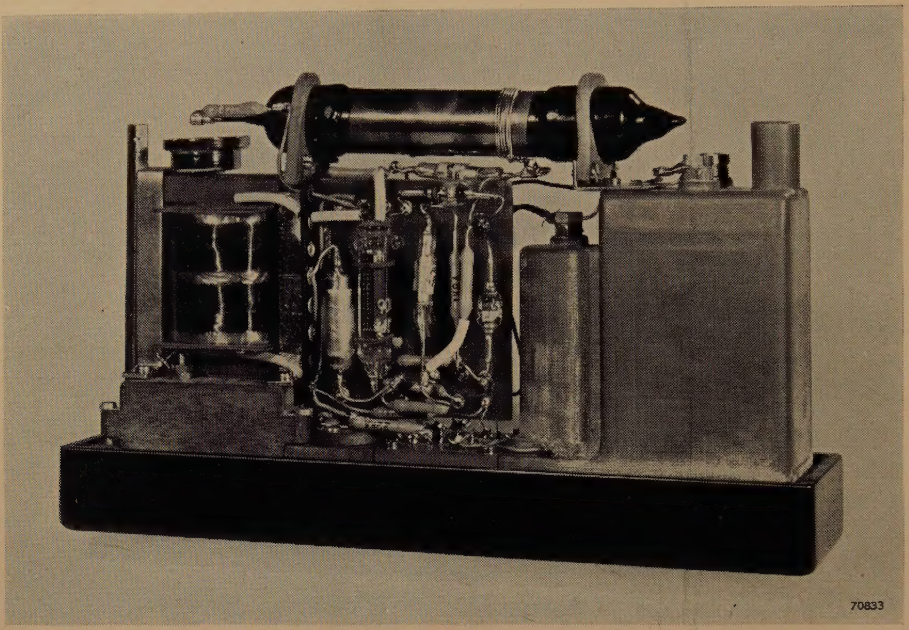

Fig. 10. The instrument with cover removed. At the top will be seen the counter tube. At left the transformer, at right the filament and H. T. batteries.

Summary. In the design of small portable Geiger counter instruments a low current consumption is of considerable importance; the smaller types of dry battery can then be used. This is achieved very successfully if the voltage for the counter tube is supplied by a blocking oscillator delivering voltage pulses which are employed to charge a capacitor in conjunction with a diode. The blocking oscillator functions only when a count in the Geiger tube withdraws some of the charge from the capacitor. An instrument has been constructed along these lines, measuring only $17 \times 4 \times 10$ cm (approx. $7'' \times 2'' \times 4''$) and weighing not more than 520 grammes (approx, 18 ozs.), the batteries being of the kind employed in hearing-aids. Some of the requirements to be met by the circuit components are discussed, and the article also describes the circuit associated with the measuring instrument, which is calibrated in terms of the number of ionizing particles reaching the counter tube. The circuit is specially designed to ensure low battery drain.